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PREFACE

B.R.C.N. 5422 (1) and (2), Radio Fundamentals (Electrical) and (Electronic) has been prepared as a textbook for use at H.M.C. Electrical School and H.M.C. Communications School.

B.R.C.N. 5422 (1) contains basic electrical theory essential to the understanding of radio and radar circuits. It has been written to provide a comprehensive review and extension of fundamental electrical principles, with their application to electronics considered of first importance.

B.R.C.N. 5422 (2) contains a study of the electronic circuits fundamental to radio and line communications, direction-finding and radar.

Work on the manuscript was originally undertaken by some thirty-three instructional officers at H.M.C. Signal School during the Second World War. The text has been revised and completed by the undersigned former members of this staff. The manuscript is based upon notes compiled by the authors while training communicators and technicians and while taking courses. Consequently the material has been drawn from a great many excellent sources and has been adapted and presented in such a way as to suit best the needs of the R.C.N. To list these sources and give credit where credit is due would be a most difficult task since knowledge and experience, in any field, are based upon facts derived from many diverse sources.

In addition to the undersigned, Lieutenant (L) J. H. Burwell, R.C.N. (R) (Retired), has helped, during two summers, in the preparation of B.R.C.N. 5422 (2).

The original manuscripts for Chapter XVIII Line Communications and Chapter XIX Teletype were written by staff officers of H.M.C. Electrical School and edited by the authors.

B.R.C.N. 5422 (1) and (2) has been produced with the co-operation of H.M.C. Electrical School and H.M.C. Communications School. The staffs of these schools have read the manuscript and have made many valuable suggestions and criticisms. In addition, B.R.C.N. 5422 (2) was used in mimeograph form in the training of two classes of technicians at H.M.C. Electrical School. Suggested revisions and extensions have been made and the text has been accepted for use in training communicators and technicians.

The authors wish to thank the Director of Naval Communications, the Staffs of H.M.C. Electrical School and H.M.C. Communications School, the Director of Publications and Printing and the Commanding Officer of H.M.C.S. "York" for their efforts and co-operation in facilitating the work on this project; and in particular the Electrical Engineer-in-Chief and his staff whose efforts have made this work possible.

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H.M.C.S. "York"
August, 1952.

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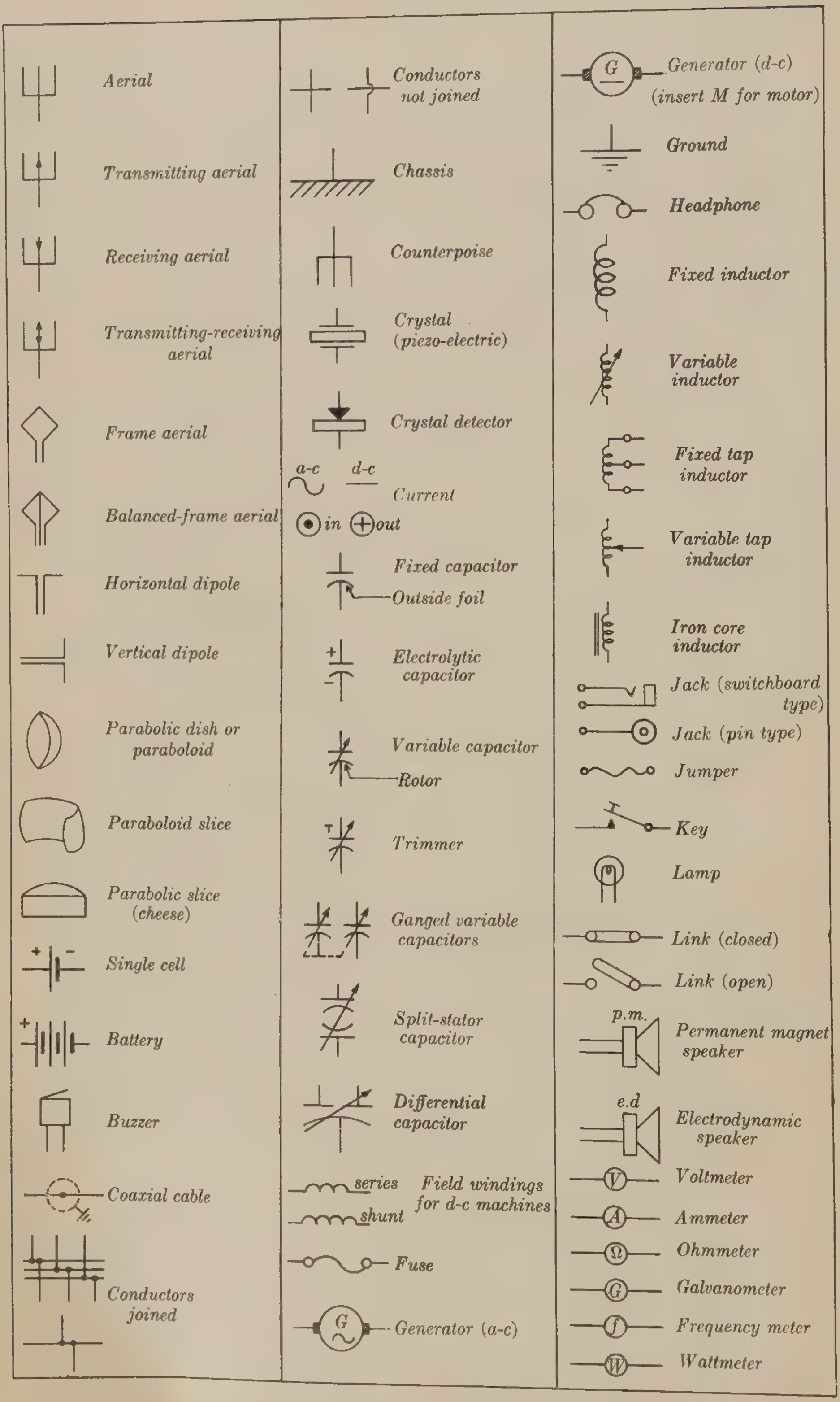
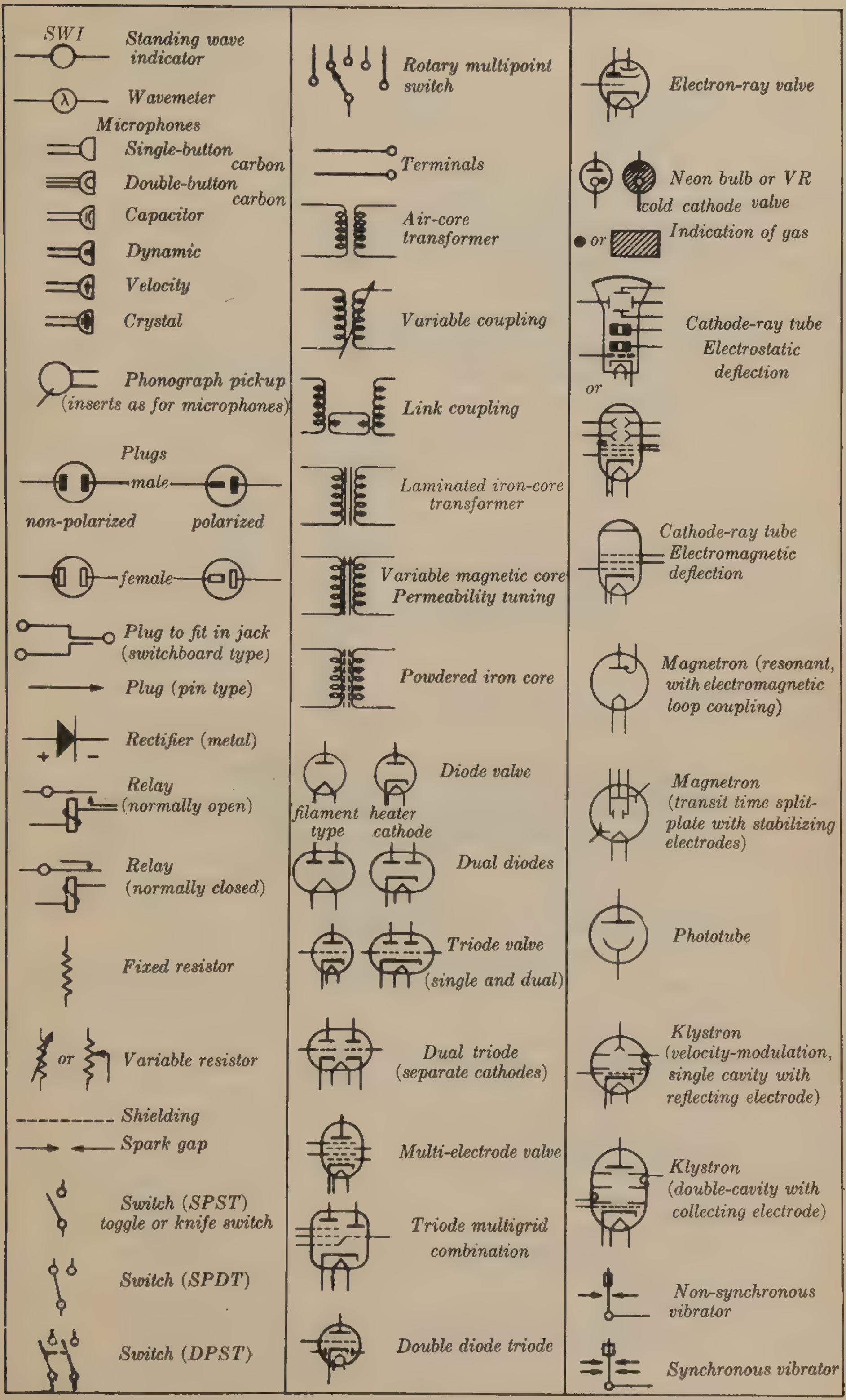

Aerial
Transmitting aerial
Receiving aerial
Transmitting-receiving aerial
Frame aerial
Balanced-frame aerial
Horizontal dipole
Vertical dipole
Parabolic dish or paraboloid
Paraboloid slice
Parabolic slice (cheese)
+
-
Single cell
+
Battery
Buzzer
Coaxial cable
Conductors joined
Conductors not joined
Chassis
Counterpoise
Crystal (piezo-electric)
Crystal detector
a-c
d-c
Current
in
out
Fixed capacitor
Outside foil
+
-
Electrolytic capacitor
Variable capacitor
Rotor
Trimmer
Ganged variable capacitors
Split-stator capacitor
Differential capacitor
series
shunt
Field windings for d-c machines
Fuse
G
Generator (a-c)
G
Generator (d-c) (insert M for motor)
Ground
Headphone
Fixed inductor
Variable inductor
Fixed tap inductor
Variable tap inductor
Iron core inductor
Jack (switchboard type)
Jack (pin type)
Jumper
Key
Lamp
Link (closed)
Link (open)
p.m.
Permanent magnet speaker
e.d
Electrodynamic speaker
V
Voltmeter
A
Ammeter
Ω
Ohmmeter
G
Galvanometer
f
Frequency meter
W
Wattmeter


SWI
Standing wave indicator
λ
Wavemeter
Microphones
Single-button carbon
Double-button carbon
Capacitor
Dynamic
Velocity
Crystal
Phonograph pickup (inserts as for microphones)
Plugs
male
non-polarized
polarized
female
Plug to fit in jack (switchboard type)
Plug (pin type)
Rectifier (metal)
+
−
Relay (normally open)
Relay (normally closed)
Fixed resistor
or
Variable resistor
Shielding
Spark gap
Switch (SPST) toggle or knife switch
Switch (SPDT)
Switch (DPST)
Rotary multipoint switch
Terminals
Air-core transformer
Variable coupling
Link coupling
Laminated iron-core transformer
Variable magnetic core Permeability tuning
Powdered iron core
Diode valve
filament type
heater cathode
Dual diodes
Triode valve (single and dual)
Dual triode (separate cathodes)
Multi-electrode valve
Triode multigrid combination
Double diode triode
Electron-ray valve
Neon bulb or VR cold cathode valve
or
Indication of gas
Cathode-ray tube Electrostatic deflection
or
Cathode-ray tube Electromagnetic deflection
Magnetron (resonant, with electromagnetic loop coupling)
Magnetron (transit time split-plate with stabilizing electrodes)
Phototube
Klystron (velocity-modulation, single cavity with reflecting electrode)
Klystron (double-cavity with collecting electrode)
Non-synchronous vibrator
Synchronous vibrator

LETTER SYMBOLS AND ABBREVIATIONS

A	Ampere
A	Amplification (Voltage gain)
A	Amplitude
A_c	Amplitude of carrier wave
A_m	Amplitude of modulating wave
a	Cross-sectional area
a-c	Alternating current
a-f	Audio-frequency
a.f.c.	Automatic frequency control
a.g.c.	Automatic gain control
a.m.	Amplitude modulation
a.v.c.	Automatic volume control
α	An angle

B	Susceptance
B	Magnetic flux density
β	Feedback factor
b.f.o.	Beat frequency oscillator

C	Capacity; Capacitance
C_{ak}	Anode-to-cathode interelectrode capacitance
C_c	Capacitance of coupling capacitor
C_e	Effective capacitance; Equivalent capacitance
C_g	Capacitance of grid capacitor
C_{ga}	Grid-to-anode interelectrode capacitance
C_{gk}	Grid-to-cathode interelectrode capacitance
C_i	Input capacitance of a valve
C_k	Capacitance of cathode by-pass capacitor
C_n	Neutralizing capacitance
c	Velocity of propagation of electromagnetic waves in free space
C.M.	Circular-mil
C.M.F.	Circular-mil-foot
c.r.o.	Cathode-ray oscilloscope
c/s	Cycles per second
c.r.t.	Cathode-ray tube
c-w	Continuous wave

D	Electric flux density
d	Distance
d	Diameter
d.a.v.c.	Delayed automatic volume control

db	Decibel
d-c	Direct current
d-f	Direction finding
d.p.	Difference of potential
$\left.\begin{matrix}\triangle\\ \delta\end{matrix}\right\}$ Delta	Small change in a quantity

E	Electric field intensity; Electric field strength
E	Effective value of an a-c electromotive force; Value of a d-c electromotive force
E_a	Effective value of the a-c component of anode voltage of a valve
$E_{a\,max}$	Peak value of a-c component of anode voltage
E_b	Value of the d-c component of anode voltage of a valve
E_{bb}	Anode-supply voltage
E_c	Value of the d-c component of grid voltage of a valve
E_{cc}	Control-grid-bias supply voltage
E_{cc2}	Supply voltage for grid number 2
E_f	Filament voltage of a valve
E_{ff}	Filament-supply voltage
E_g	Effective value of the a-c component of control grid voltage of a valve
E_{g2}	Effective value of the a-c component of screen-grid voltage of a valve
E_{max}	Maximum value of an a-c voltage; peak value of an a-c voltage
E_p	Polarizing voltage
e	Instantaneous value of voltage
e_a	Instantaneous value of the a-c component of the anode voltage of a valve
e_b	Instantaneous value of the total anode voltage of a valve
e_c or e_{c1}	Instantaneous value of the total grid voltage of a valve
e_{c2}	Instantaneous value of screen-grid voltage of a valve
e_g	Instantaneous value of the a-c component of the control-grid voltage of a valve
e_{g2}	Instantaneous value of the a-c component of screen-grid voltage of a valve
e_i	Instantaneous value of input voltage
e_s	Signal voltage
e.m.f	Electromotive force
ε (Epsilon)	Base of natural logarithms (2.71828......)
ε	Charge on an electron
η	Anode efficiency

F	Farad
F	Force
f	Frequency
f-a	Fore and aft
f_r	Resonant frequency
f_d	Frequency deviation
f-m	Frequency-modulation

G	Conductance
G_i	Input conductance of a valve
g_m	Mutual conductance, or control-grid-to-anode transconductance of a valve
γ (Gamma)	An angle
H	Magnetic field strength
H	Henry
HP	Horsepower
$H.T.$	High tension
h_{eff}	Effective height
$h\text{-}f$	High-frequency
I	Effective value of an alternating current; value of d-c current
I_a	Effective value of the a-c component of anode current of a valve
$I_{a\,max}$	Peak value of a-c component of anode voltage
I_b	Value of the d-c component of anode current of a valve
I_{b0}	D-c no-signal component of anode current
I_{ba}	Increase in the d-c component of anode current when a signal is applied
I_c	Value of the d-c component of the grid current of a valve
I_f	Filament current of a valve
I_g	Effective value of the a-c component of the grid current of a valve
I_k	Effective value of the a-c component of cathode current of a valve
I_{g2}	Effective value of the a-c component of screen-grid current of a valve
i	Instantaneous value of current
i_a	Instantaneous value of the a-c component of the anode current of a valve
i_b	Instantaneous value of the total anode current of a valve
i_c	Instantaneous value of the total grid current of a valve
$i\text{-}f$	Intermediate-frequency
i_g	Instantaneous value of the a-c component of the control-grid current of a valve
i_{g2}	Instantaneous screen-grid current of a valve
i_k	Instantaneous cathode current of a valve
j	$\sqrt{-1}$, operator j
K	Dielectric constant
K	Kilo
k	coupling factor
L	Inductance
L_{crit}	Critical value of inductance
L_e	Equivalent or effective inductance
L_l	Effective inductance of the load
L_p	Inductance of a transformer primary
L_s	Inductance of a transformer secondary
$L.T.$	Low Tension
l	Length

l-f	Low-frequency
l.o.	Local oscillator
log	Common logarithm; Logarithm to the base 10
λ (Lambda)	Wavelength

M	Mutual inductance
M.F.	Mil-foot
M	Mega
m	Modulation factor
m-c-w	Modulated continuous wave
m_f	Frequency modulation index; deviation ratio
msec.	Millisecond
m-f	Medium-frequency
m.m.f.	Magnetomotive force
m.o.	Master oscillator
m.o.p.a.	Master oscillator power amplifier
μ (Mu)	Amplification factor of a valve; Permeability

N	Number of turns
N_p	Number of transformer primary turns
N_s	Number of transformer secondary turns

O	Operating point
Ω (omega)	Ohm
ω (omega)	Angular velocity
ω_r	Resonant angular velocity

P	Power; true power
$P_{a\text{-}c}$	A-c power
P_i	Input power
P_o	Output power
p	Instantaneous value of power
p.a.	Power amplifier
p.d.	Potential difference
p.f.	Power factor
p.m.	Permanent magnet
p-m	Phase modulation
p-t-m	Pulse-time modulation
p.p.	Push-pull
p.s.	Power supply
p-s	Port and starboard
π (pi)	Ratio of circumference to diameter of a circle (3.14159. .)
ϕ (Phi)	Phase angle of current
Φ (Phi)	Total magnetic flux

Q	Quantity
Q	Quantity of charge on a capacitor
Q	Ratio of reactance to resistance for an inductor
Q	Figure of merit of a tuned circuit
q	Instantaneous electric charge
R	Resistance
R_a	Direct-current resistance of anode circuit
R_{aa}	Anode-to-anode resistance
R_{crit}	Critical value of resistance
R_e	Effective resistance; Equivalent resistance
R_g	Direct-current resistance of control-grid circuit
R_k	Direct-current resistance of cathode circuit
R_L	Load resistor
R_l	Control-grid current limiting resistor
R_p	Resistance of transformer primary
R_s	Resistance of transformer secondary
r	Radius
r	Internal resistance of a source of e.m.f.
r_a	Dynamic anode resistance of a valve; A-c resistance of a valve
r_b	Direct-current resistance of a valve
r_i	Input resistance of a valve to a-c
r-f	Radio-frequency
r-f-c	Radio-frequency choke
r.m.s.	Root mean square
r.p.m. or *r/m*	Revolutions per minute
r-t	Radio telephony
ρ (Rho)	Resistivity; specific resistance
S	Sensitivity or signal strength
s-f	Supersonic frequency
s.g.	Specific gravity
T	Absolute temperature
T	Period of sinusoidal wave
t	Time
t	Turns ratio
t-r-f	Tuned-radio-frequency
θ (Theta)	Phase angle
u-h-f	Ultra-high-frequency
V	Volt; Potential difference; Voltage drop; Potential
v	Velocity
v-h-f	Very-high-frequency
v-u	Volume unit

W	Work or energy
W	Watt
WH	Watt-hour
$W.V.$	Voltage rating or working voltage
w-t	Wireless telegraphy
X	Reactance
X_c	Capacitive reactance
X_e	Effective reactance; Equivalent reactance
X_i	Input reactance
X_L	Inductive reactance
X_o	Output reactance
Y	Admittance
Y_i	Input admittance
Y_o	Output admittance
Z	Impedance
Z_a	Impedance of anode circuit
Z_e	Effective impedance; Equivalent impedance
Z_i	Input impedance
Z_g	Impedance of control-grid circuit
Z_k	Impedance of cathode circuit
Z_o	Output impedance; Characteristic impedance

FIRST-AID TREATMENT FOR ELECTRIC SHOCK

A. Safety First

Regard all electrical equipments as dangerous irrespective of voltage. Exercise great care in handling as it is often a minor shock which causes an operator to contact more dangerous circuits with serious results. When handling equipment, avoid broad contacts such as are made by standing on a metal deck or in water. Stand on a piece of good insulating material when necessary to make adjustments with equipment in operation. In addition, make adjustments with one hand only wherever possible.

Dangerous contacts may result through lessened resistance when the skin and clothing are wet with perspiration. Contact with damp metal surfaces such as decks, bulkheads, guns and machinery, may allow current to ground through the moist skin and body. Electric shock is due to current passing through the body irrespective of voltage. An 110-volt source has caused death. Current passing through the body in the region of the heart is especially dangerous. In using electric breast drills avoid the possibility of a ground.

Usually electric shock does not kill instantly. Life can be saved even though breathing has stopped.

B. Instructions For Resuscitation

A step-by-step analysis of the Holger-Nielson
Method of Artificial Respiration

Follow these instructions even if the victim appears dead.

I. Free the victim from the circuit immediately.

1. Quickly release victim from current, being careful to avoid receiving a shock. Use any dry non-conductor (rubber gloves, clothing, wood, rope, etc.) to remove either victim or conductor. Beware of using metal or any moist material. If necessary shut off current.

2. As soon as victim is clear of conductor rapidly feel with your finger in his mouth and throat and remove any foreign body (tobacco, false teeth, etc.). If mouth is tight shut, pay no more attention to it until later; these aspects can be cared for when placing the subject into position or shortly thereafter, between cycles. Do not stop to loosen patient's clothing, but immediately begin actual resuscitation. Every moment of delay is serious. Proceed as follows.

II. Instantly attend to victim's breathing.

POSITION OF THE SUBJECT.

1. Place the subject in the facedown, prone position. Bend his elbows and place his hands one upon the other. Turn his face to one side, placing his cheek upon his hands. (See Fig. 1).

POSITION OF THE OPERATOR.

2. Kneel on either the right or left knee at the head of the subject, facing him. Place the knee at the side of the subject's head, close to his forearm. Place your opposite foot near his elbow if it is more comfortable, kneel on both knees, one on either side of the subject's head. Place your hands upon the flat of the subject's back in such a way that the heels lie just below a line running between the armpits. With the tips of the thumbs just touching, spread the fingers downward and outward. (See Fig. 2).

COMPRESSION PHASE.

3. Rock forward until your arms are approximately vertical, and allow the weight of the upper part of your body to exert slow, steady, even pressure downward upon the hands. This forces air out of the lungs of the subject. Your elbows should be kept straight, and the pressure exerted almost directly downward on the back. (See Fig. 3).

EXPANSION PHASE.

4. Release the pressure, avoiding a final thrust, and commence to rock slowly backward. Place your hands upon the subject's arms just above his elbows, and draw his arms upward and toward you. Apply just enough lift to feel resistance and tension at the subject's arms shoulders. Do not bend your elbows and as you rock backward, the subject's arms will be drawn toward you. Then drop the arms gently to the ground. This completes the full cycle. The arm-lift expands the chest by pulling on the chest muscles, arching the back, and relieving the weight on the chest. (See Fig. 4).

5. The cycle should be repeated 12 times per minute at a steady, uniform rate. The compression and expansion phases should occupy about equal time; the release periods being of minimum duration. A smooth rhythm in performing artificial respiration is desirable, but split-second timing is not essential.

6. Continue artificial respiration without interruption until natural breathing is restored; if necessary, four hours or longer, or until a physician declares the patient is dead.

7. As soon as this artificial respiration has been started and while it is being continued, an assistant should loosen any tight clothing about the patient's neck, chest or waist. *KEEP THE PATIENT WARM.* Do not give any liquids whatever by mouth until the patient is fully conscious.

8. To avoid strain on the heart when the patient revives, he should be kept lying down and not allowed to stand or sit up. If the doctor has not arrived by the time the patient has revived, he should be given some stimulant, such as one teaspoonful of aromatic spirits of ammonia in a small glass of water or a hot drink of coffee or tea, etc. The patient should be kept warm.

9. Resuscitation should be carried on at the nearest possible point to where the patient received his injuries. He should not be moved from this point until he is breathing normally of his own volition, and then moved only in a lying position. Should it be necessary, due to extreme weather conditions, etc., to move the patient before he is breathing normally, resuscitation should be carried on during the time that he is being moved.

10. A brief return of natural respiration is not a certain indication for stopping the resuscitation. Not infrequently the patient after a temporary recovery of respiration, stops breathing again. The patient must be watched and if natural breathing stops, artificial respiration should be resumed at once.

11. In carrying out resuscitation, it may be necessary to change the operator. This change must be made without losing the rhythm of respiration. By this procedure no confusion results at the time of change of operator and a regular rhythm is kept up.

III. Send for the Medical Officer or nearest physician.

If alone with victim, do not neglect immediate and continued resuscitation in order to call the M.O. Start at once—the first few minutes are valuable. If other persons are present, send one of them for a doctor without a moment's delay.

The ordinary and general tests for death should not be accepted and any doctor should make several very careful and final examinations and be sure specific evidence is present before pronouncing the patient dead.

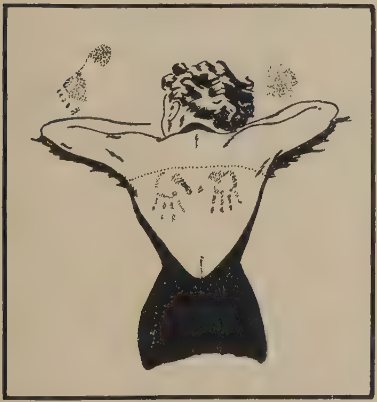


FIGURE 1

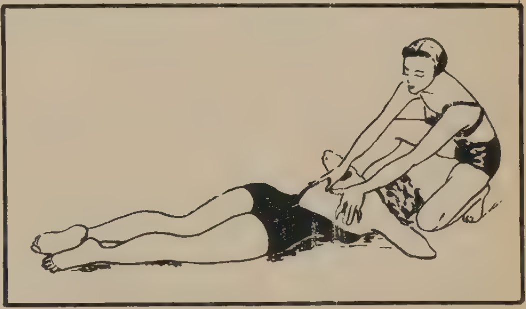


FIGURE 2

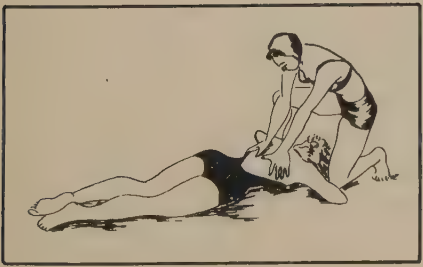


FIGURE 3

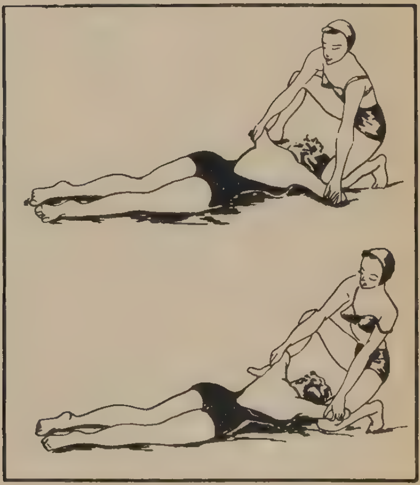


FIGURE 4

CHAPTER I

VALVES

The *thermionic valve* or *thermionic tube* is one of the most important of all the various components used in modern radio systems. A valve consists of a glass or metal container in which two or more *electrodes* are mounted on a suitable supporting structure. Some valves are highly evacuated while others contain an inert gas at low pressure. Electrical connections are brought to pins or terminals on the outside of the container. Electrons emitted from one of the electrodes form an *electron stream.* The formation of this electron stream inside the valve is therefore the first consideration.

1.1 Electronic emission and thermionic emission. The supply of free electrons for the electron stream is formed inside the valve by the emission of electrons from one of the electrodes called the *cathode.* The process by which electrons are set free is called *electronic emission.*

In Chapter II, Vol. 1, it is explained that conductors are made up of atoms that give up orbital electrons easily and hence have a great many free electrons which are in continual random motion within the substance. If the conductor is heated, the free electrons gain velocity. At high temperatures, when the material begins to glow, some free electrons acquire a velocity sufficiently high to permit them to escape by penetrating the surface of the conductor and these form a cloud of electrons around the conductor. This is one method of electronic emission and in this case since the required energy for emission is supplied to the free electrons in the form of heat, the emission is called *thermionic emission.* The substance from which the electrons escape is called the *emitter* or the cathode. The rate at which electrons escape from the emitter surface increases

(i) as the temperature increases
(ii) as the emitting surface area increases
(iii) as the surrounding pressure becomes less.

Thus, if an emitter of relatively large surface area, made of a substance rich in free electrons, is mounted and sealed in an evacuated glass or metal tube, it becomes, when heated to a high temperature, a practical source of electrons for the operation of electronic valves. However the high temperatures required to produce satisfactory emission limit the number of substances that may be used as emitters to a very few. The most commonly used emitting materials are tungsten, thoriated tungsten and oxide-coated emitters.

Tungsten, when raised to white heat, gives good electronic emission and is referred to as a *bright emitter.* Emitters of this material require heating to a very high temperature but they are rugged and have a long life; hence they are used in high-power transmitting valves.

Thoriated tungsten consists of tungsten in which a small amount of thorium oxide has been incorporated. This material gives a satisfactory electron emission at a very much lower temperature than pure tungsten, operating at yellow or orange heat. Emitters of this material are used in medium- and low-power transmitting valves and in the larger receiving valves.

Oxide-coated emitters consist of a wire or ribbon of nickel or platinum alloy

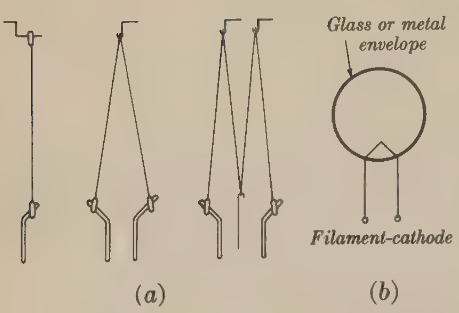


FIG.1-1 (a) Directly-heated filament cathodes (b) Circuit symbol of a valve with directly-heated cathode.

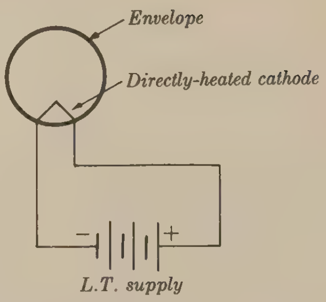


FIG.1-2 L.T. supply connected to heat the valve filament.

with its surface coated with a mixture of the oxides of barium and strontium. From this coating large numbers of electrons are emitted at relatively low temperatures corresponding to a dull red heat, and this type of emitter is therefore termed a *dull emitter*. Many of the medium-size valves and practically all of the receiving valves use oxide-coated cathodes.

1.2 Heating the emitter. The emitters of valves are heated electrically. One method is to construct the emitter in the form of a filament, Fig. 1-1 (a), and connect it across a battery or source of alternating current. Since such an emitter is a good conductor its resistance is low, the current in it is large and the heating effect (I^2R) is also large. In order to supply the desired high current only a low-voltage source of high capacity is required. Such a low voltage-high current source is called a *Low Tension* (L.T.) or *"A" supply*. For example, some transmitting valves require an L.T. supply capable of delivering 4.5 amperes at 10 volts, while some receiving valves require an L.T. supply which will provide 0.15 amperes at 6.3 volts. Thus by means of such a supply the emitter may be maintained at the required temperature for satisfactory emission. If the emitter itself conducts the heating current, as in this case, the cathode is said to be *directly heated* and it is referred to as a *directly-heated cathode*. This type of cathode may be of tungsten, thoriated tungsten or of the oxide-coated type. The circuit symbol for a valve with a directly-heated cathode is shown in Fig. 1-1 (b) and the circuit arrangement with L.T. supply connected is shown in Fig. 1-2.

A second method of heating is to construct the emitter in the form of a hollow cylinder and the heating is accomplished by mounting a tungsten heater wire or filament inside the cylinder. The heater only is connected across an L.T. supply and the cylindrical emitter is *heated indirectly* from the heat generated by the heater wire. This type of emitter is said to be *indirectly heated* and it is referred to as an *indirectly-heated cathode*. Typical indirectly-heated cathodes are shown in Fig. 1-3 (a), the valve circuit symbol in Fig. 1-3 (b), and valve arrangement in Fig. 1-4.

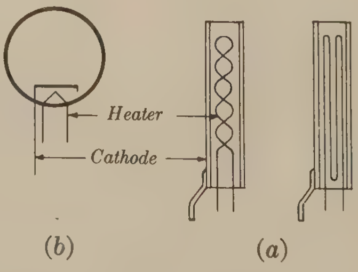


FIG.1-3 (a) Indirectly-heated cathodes (b) Circuit symbol of valve with indirectly-heated cathode.

If the customary 110-volt commercial a-c power mains is to be used to provide the L.T. supply, a step-down transformer, called a *filament transformer* is used to step down the line voltage to the low-voltage rating of the filament of the valve. Such a circuit arrangement is shown in Fig. 1-5.

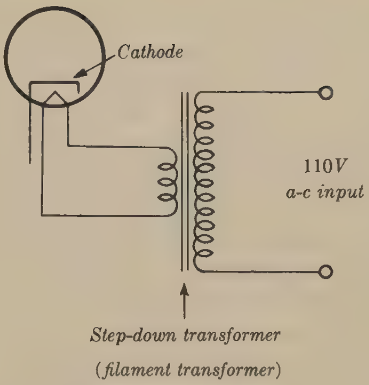


FIG. 1-5 Arrangement for a-c heating of the emitter.

1.3 The electron stream. If a positively charged body is placed near an electron emitter in an evacuated envelope, Fig. 1-6 (a), it attracts and collects the emitted electrons (unlike charges attract). The collected electrons tend ultimately to neutralize the positive charge of this body. However, if the electrons which are collected are removed as quickly as they arrive, the positive charge of the *collector electrode* can be maintained. The removal of the collected electrons may be accomplished by connecting a source of constant d-c voltage between the positively charged collector electrode and the emitter as shown in Fig. 1-6 (b). Thus the electrons emitted by the cathode are attracted to the anode, where they move from atom to atom as free electrons within the metal. The electron flow continues through the battery back to the cathode to replenish its supply. This flow of electrons constitutes a flow of current. The two-electrode valve is known as a *diode*. The collector electrode of the valve is referred to as the *anode* or *plate* and it is usually constructed in the form of an open-ended cylinder or box of sheet metal mounted to surround the cathode. When it is necessary

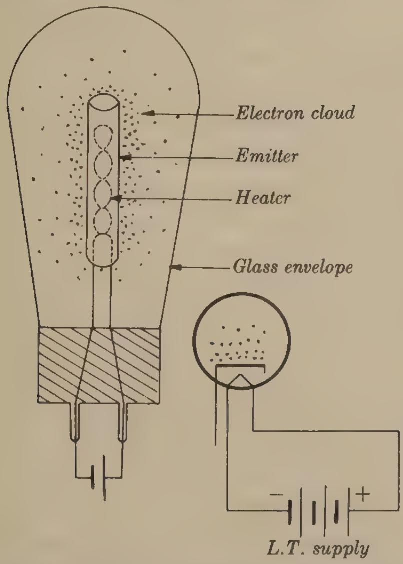


FIG. 1-4 L.T. supply arrangement for indirectly-heated cathode valve.

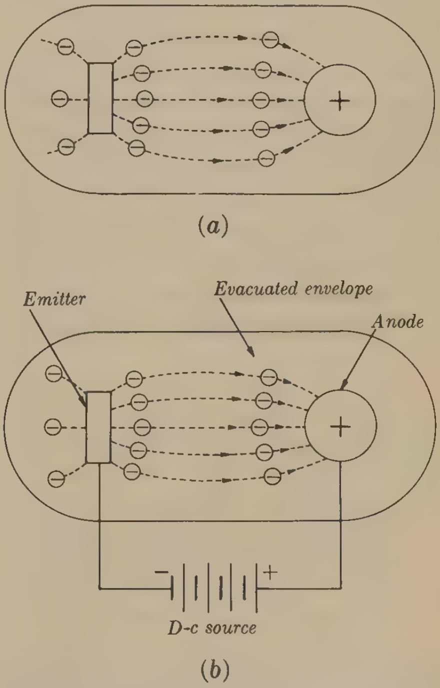
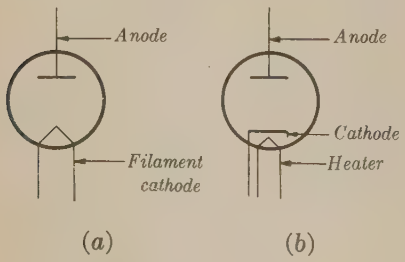


FIG. 1-6 Function of collector electrode (anode).

to maintain the anode at a steady positive potential, a d-c supply, known as the *High Tension* (H.T.) or *"B" supply* is used. In order to supply the desired current the d-c supply is generally one of high voltage. Since the current is small the H.T. supply need only have a low capacity. To maintain the anode at a positive potential, it is connected to the positive terminal of the supply while the cathode is connected to the negative terminal. In the case of transmitters and receivers, the H.T. supply voltages are often of the order of 2000 volts and 350 volts respectively.

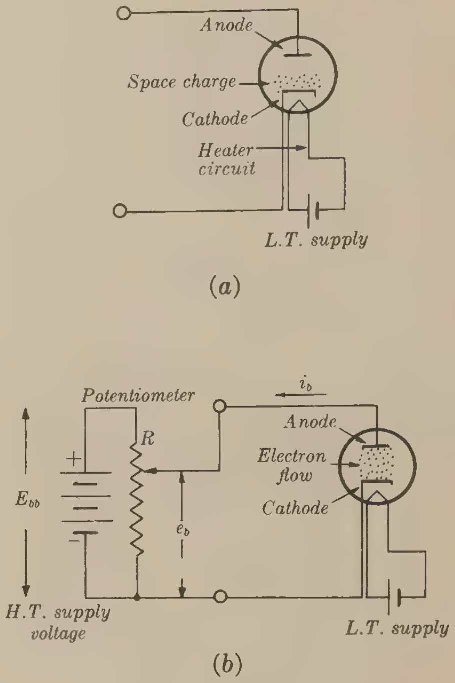
1.4 The diode. A diode valve has two electrodes, a cathode and an anode. Fig. 1-7 (a) and (b) show the circuit symbols used for a diode with a directly-heated cathode and an indirectly-heated one respectively.

A more complete understanding of the operation of the diode valve may be obtained by a consideration of the circuits shown in Fig. 1-8. In Fig. 1-8 (a) when the L.T. supply is connected to the heater, as shown, the cathode emits electrons as previously explained. In the absence of a potential on the anode these electrons form a cloud around the cathode, this cloud being called the *space charge*. This space charge tends to repel electrons, which are just being emitted, back into the cathode. As a result, the space charge increases in size until the same number of electrons are being repelled into the cathode by the space charge as are being emitted by the cathode.



(a) (b)

Fig. 1-7 Circuit symbols for diode valve (a) with directly-heated cathode (b) with indirectly-heated cathode.



(a)

(b)

Fig. 1-8 Circuit for studying operation of the diode.

If an H.T. supply voltage represented by the abbreviation E_{bb} is now connected across anode and cathode via the voltage divider R, as shown in Fig. 1-8 (b), the anode is made positive to the cathode. When the voltage divider tapping point is set to make the anode only slightly positive with respect to the cathode some electrons from the space charge are attracted to the anode and return through the external circuit of the H.T. supply to the cathode. This flow of electrons is known as the *anode current* (or *plate current*) of the valve and the path it takes is referred to as the *anode circuit* (or *plate circuit*). The value of the anode current flowing at any instant is represented by the abbreviation i_b and it may be measured by inserting a milliammeter in the anode circuit. The p.d. applied between anode and cathode is referred

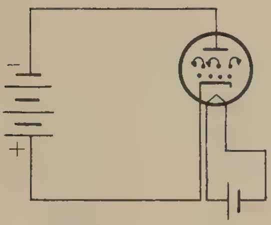


FIG.1-9 Diode with anode negative to the cathode.

to as the *anode voltage* (or *plate voltage*) of the valve and its value at any instant is represented by the abbreviation e_b.

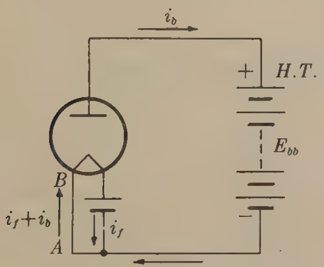
When the anode voltage is made more positive, there is an increase in the number of electrons flowing to the anode, that is i_b increases with an increase in e_b. A limit is reached when an increase in e_b causes no further increase in i_b due to the fact that all the available electrons emitted by the cathode are being collected by the anode. The value of anode current under these conditions is known as the *saturation current* and the condition is called *anode saturation*. If the cathode temperature is less the electron emission is less. Hence the anode voltage at which saturation sets in and the value of the saturation current are less.

It should be noted that if the anode is made negative with respect to the cathode (Fig. 1-9) the space charge electrons are repelled by the anode (like charges repel) and no anode current flows. Thus a diode permits an electron flow from cathode to anode but not from anode to cathode. This unidirectional property of the diode is of first importance in the application of this valve in radio circuits.

In the circuit of Fig. 1-10, the cathode is of the directly-heated type and the anode current (i_b) and the filament current (i_f) flow in the same lead *AB*. If an a-c filament voltage is used with a directly-heated cathode, emission may vary to a slight extent in step with the a-c. This condition is undesirable and as a result indirectly-heated cathodes are usually employed when a-c is used as the L.T. supply.

1.5 The characteristic curve of a diode. A continuous picture of the variation in the anode current of a diode as the anode voltage is gradually increased may be obtained by measuring corresponding $e_b - i_b$ values and plotting them to obtain a graphical representation of their relationship.

The curve which shows this $e_b - i_b$ relationship is called the $e_b - i_b$ *curve* or the *anode characteristic* of the diode. The corresponding $e_b - i_b$ values may be obtained by using the circuit of Fig. 1-11. With the aid of potentiometers both the filament and anode voltages may be varied from zero up to the rated values of L.T. and H.T. specified by the manufacturer for the valve used. The voltmeters V_1 and V_2 measure the anode and filament voltages respectively while the milliammeter *mA* measures the anode current. With a constant filament



Arrows show direction of electron flow

FIG.1-10 Circuit of diode with directly-heated cathode.

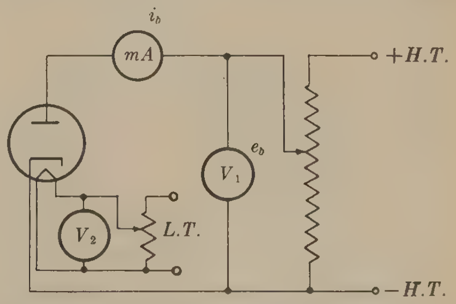


FIG.1-11 Circuit for obtaining data to plot the characteristic curves of a diode.

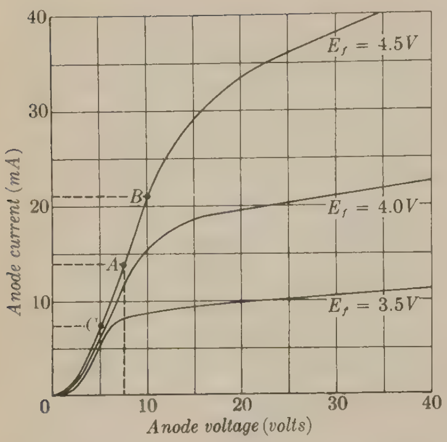


FIG. 1-12 Diode static anode characteristics.

voltage the anode voltage is varied in steps and the corresponding values of anode current are noted. The values of anode voltage are used as abscissae and the values of anode current as ordinates to plot the characteristic curve. The curve is obtained by drawing a smooth line through the plotted points (Fig. 1-12).

In the graph of Fig. 1-12 the saturation current is approximately 12mA when the filament voltage is 3.5V. The nearly horizontal portion of the curve represents the saturation condition. At saturation the anode current is independent of the anode voltage. If the filament voltage is increased to 4.0V, more electrons are emitted by the cathode. In this case the anode current for corresponding values of anode voltage is greater than the anode current when the filament is maintained at 3.5V. Hence, while the general shape of the characteristic is unchanged, the saturation current is about 24mA. If the filament voltage is increased still further to 4.5V, saturation does not set in at anode voltages shown in Fig. 1-12. In obtaining the $e_b - i_b$ values to draw these curves it is necessary to operate the filament at a voltage considerably lower than its rated value so that the anode current does not exceed its maximum rated value.

The curves shown in Fig. 1-12 are typical of diodes with tungsten or thoriated-tungsten cathodes. The comparative anode characteristics for diodes which have oxide-coated cathodes is shown in Fig. 1-13. From this curve it may be seen that the anode current never reaches a value of complete saturation. This results from the irregularities in the coating. Due to these irregularities small pockets are produced which retain tiny space charges since they are somewhat protected from the anode potential. Hence space-charge electrons continue to be drawn from these pockets after most of the principal space charge has disappeared and so the anode current continues to increase slightly as shown in the graph.

The value of the ratio $\frac{\text{anode voltage}}{\text{anode current}}$ is called the *equivalent d-c resistance* of the diode. For example, at the point A on the curve for which the filament voltage is 4.5V, the anode voltage is 7.5V and the anode current is 14.0mA. The equivalent d-c resistance at A is $\frac{7.5}{14.0\times10^{-3}}$ ohms or 536 ohms approximately. Similarly at B and C the equivalent d-c resistance is 476 ohms and 714 ohms respectively. The equivalent d-c resistance of the diode is a variable depending on the operating conditions.

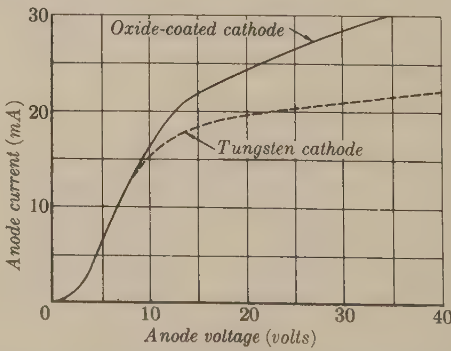


FIG. 1-13 Comparative anode characteristics of diodes with oxide-coated cathode and tungsten cathode.

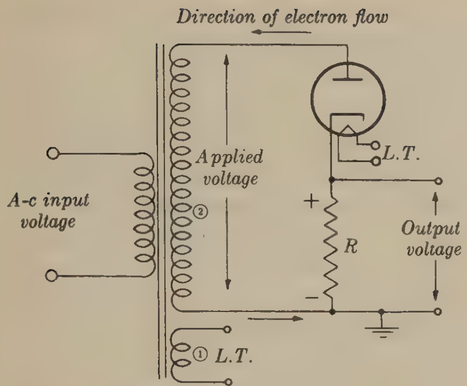


Fig.1-14 Diode rectifier circuit.

A more important property of the valve is the *a-c anode resistance* (r_a) which is the ratio of the change in anode voltage to the resulting change produced in anode current. If a steady anode voltage of 7.5V is applied, point A in Fig. 1-12, and then an alternating voltage of 2.5V is superimposed, the resulting change in valve current has a peak value of 7mA, C to B. Since a voltage change of 5V causes a current change of 14mA the a-c anode resistance is $\frac{5}{14.0\times10^{-3}}$ ohms or 357 ohms approximately. Over the straight line portion of a diode characteristic the a-c anode resistance is constant. The steeper the slope of the characteristic the less is the a-c resistance.

1.6 The diode as a rectifier. The ability of a diode to permit current to flow through it in one direction only makes it possible to convert alternating current into direct current. This is one of the most important uses of a diode in radio circuits. When a diode is used for this purpose it is known as a *rectifier*.

The simplest form of diode rectifier circuit is shown in Fig. 1-14. The 110V a-c main is connected across the primary of a power transformer which has two secondary windings. Winding (1) steps down the 110-volt input to the rated heater voltage of the diode and when connected as shown provides the necessary current to heat the cathode. Winding (2) has a step-up ratio and the voltage developed across this winding is applied across anode and cathode of the valve through the resistor R. On the positive half cycles the anode is positive with respect to the cathode and the diode conducts electrons from cathode to anode, through winding (2) and resistor R back to the cathode. This current produces an IR drop across R. On the negative half cycles the anode is negative to the cathode and there is no electron flow. As the current through the diode and through the load R flows in one direction only, it is direct current. This direct current shown in Fig. 1-15 is quite different from a steady direct current. It increases from zero to a maximum value and then decreases to zero again during each positive half cycle of alternating current. Further, this current is zero during the negative half cycles. Thus the current through the valve and resistor comes in pulses and as a result it is referred to as a *pulsating direct current.* The output voltage appearing across R is also a pulsating voltage. To convert this pulsating direct current into a steady direct current and so provide a steady output voltage across R, the variations

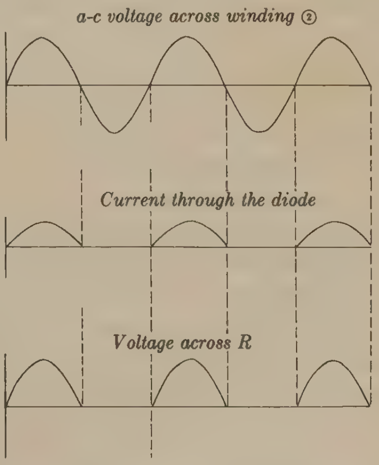


Fig.1-15 Diode rectifier circuit waveforms.

in current must be removed. The pulsation in current is called the *ripple* in the current and it can be removed by a *filter circuit.* The process of filtering out the pulsations in current to provide a steady direct current is discussed in Chapter III. A circuit which conducts current only on one half cycle of the a-c input is called a *half-wave rectifier.*

1.7 Physical construction of the cathode. Since the cathode is the heart of the vacuum tube a more detailed study of the physical construction, efficiency, advantages and disadvantages of the two main types of cathodes is desirable.

A. The Directly-Heated Cathode

If the emitter itself conducts the heating current, the cathode is a directly-heated cathode or *filament cathode.* To allow for the expansion of the wire a spring suspension is employed in the construction as shown in Fig. 1-1 (a). This type of cathode may be made of tungsten, thoriated tungsten, or it may be the oxide-coated type.

Because of its small surface area, the directly-heated cathode can be maintained at its operating temperature with a relatively small amount of heat wastage and so with relatively small power consumption. It is therefore useful in battery-operated equipment. This type is also used in large transmitting valves in order to keep power consumption to a minimum. Because the heater voltage is applied directly to the cathode all sections of this cathode are not at a uniform potential. For example, if the filament voltage is $2.5V$, the voltage difference between the anode and the negative end of the filament is $2.5V$ greater than that between the anode and the positive end of the filament. Hence electrons are not drawn to the anode in equal amounts from each end of the filament. As a result, the negative end of the filament may fail in emission sooner than the positive end.

Although the source of heater voltage may be a-c or d-c, the use of a-c for directly-heated cathodes is undesirable since the heater current alternately adds to and subtracts from the anode current in the cathode. This fluctuation results in an unwanted hum in the loudspeaker or earphones of receivers and amplifiers. Therefore d-c is used wherever possible. However, when it is necessary to use a-c with a directly-heated cathode, a valve requiring a low filament voltage should be selected to reduce this effect.

B. The Indirectly-Heated Cathode

A cathode in which the emitter is heated by a nearby current-carrying conductor or filament is called an indirectly-heated cathode or *heater cathode.* The heater wire or filament is usually made of tungsten or a tungsten-molybdenum alloy. The emitter consists of an oxide-coated metal cylinder which closely surrounds the filament. The two components are electrically insulated by a very thin coating of insulation on the heater wire.

The indirectly-heated cathode possesses the following advantages over the directly-heated cathode.

(i) The heater voltage may be a-c as well as d-c since the heater current does not pass through the emitter and hence has no undesirable effect on the anode current.

(ii) Since all sections of the emitter are at a uniform potential electrons are drawn in equal amounts from all its parts.

(iii) The emitter can be made rigid and hence the spacing between it and the other electrodes can be made smaller.

(iv) The tungsten heater can carry large heater currents with little danger of burning out.

The main disadvantage of the indirectly-heated type of cathode is that more heat must be developed and hence more power is consumed in maintaining both filament and emitter at the operating temperature. In practice the use of the indirectly-heated cathode is restricted to low-power transmitting and receiving valves.

C. Pressure Conditions for Emission

If a cathode mounted in air at normal pressure were heated to the high temperature required for thermionic emission, the oxygen present would cause the cathode material to burn.

A method of overcoming this is to mount the electrodes in an evacuated envelope. Valves which employ a high vacuum are called *hard valves*. To ensure a high vacuum many precautions are taken in manufacture to remove all gas from walls and electrodes. However, most cathode materials retain slight amounts of gas which may be given up from the surface pores under operating temperatures. Such expelled gases are called *occluded gases* and the effect is called *occlusion* of gases. Collisions of electrons passing from cathode to anode with atoms of occluded gases have two undesirable effects.

(i) The electrons are slowed down.

(ii) Electrons may be knocked from their orbits in the gas atoms, thus ionizing the gas.

As a result the atom or molecule which has lost an electron becomes a positive ion which is attracted to the cathode. If the anode voltage is large, these positive ions may acquire considerable kinetic energy and so strike the cathode with sufficient velocity to overheat the cathode surface or strip off the coating. This effect, called *positive-ion bombardment* or *cathode bombardment*, considerably shortens the life of a cathode and interferes with normal operation of the valve.

However, in some valves, gas at very low pressure is introduced for specific purposes and ionization of the gas present forms a part of their operation. Such valves are called *soft valves* or *gas valves*. Soft valves are considered in section 1.26.

D. Emitting Materials

The rate of emission for a given cathode material depends on the emission efficiency and the durability of the material. The most efficient emitting substance is that which produces a large number of electrons at a low operating temperature. *Emission efficiency* is defined as the number of milliamperes emission from one square centimeter of emitting surface per watt of electrical power dissipated in heating that square centimeter to normal operating temperature. Emission efficiencies for commonly used materials range from 3 to $1000 mA$ per sq. cm. per watt, while operating temperatures vary from 800 to $2400°C$.

A durable cathode material should not melt or vaporize at its operating temperature; it should be able to resist positive-ion bombardment for normal operating voltages and it should have as little tendency towards gas occlusion as possible.

1. TUNGSTEN CATHODES.

Tungsten is the metal commonly used for directly-heated cathodes. A tungsten cathode is very durable under overloads or positive-ion bombardment and has a long life. Hence it is most useful for high-power transmitting valves where anode potentials are $5000 V$ or more. However, because of the high operating temperature, from 2100 to $2400°C$, its emission efficiency is low, about $3-15$ mA per sq. cm. per watt. At this temperature the metal is a dazzling white color and is therefore called a *bright emitter*. Tungsten is also used in indirectly-heated cathodes as the heater element. Its high melting temperature is an advantage here. Because of this property it is capable of producing the high temperature necessary to heat the emitting surface without melting.

2. Thoriated-Tungsten Cathodes.

A thoriated-tungsten emitter is used in indirectly-heated cathodes. It consists of tungsten in which a small amount of thorium oxide has been incorporated. Since its operating temperature, 1600°C, is much lower than that for tungsten, its emission efficiency is higher, ranging from 5 to 100mA per sq. cm. per watt. However for anode voltages over 5000 V, it is subject to damage by positive-ion bombardment. Its chief use, therefore, is in medium-power transmitting valves with anode voltages from 500 to 5000 volts.

The thorium at the surface gradually evaporates and emission may fail after extensive use. However the emission of this cathode can be renewed by passing a larger current than normal through the cathode with no voltages on the other electrodes. The above-normal temperature that is produced causes the thorium oxide to be reduced and thorium to be diffused or boiled to the surface. This new layer replaces the thorium which has evaporated from the surface.

3. Oxide-Coated Cathodes.

In this type a mixture of barium and strontium oxides, which acts as the emitter, is sprayed on a metallic base. This base may be composed of platinum, platinum alloys or alloys of nickel. For a directly-heated cathode the coating is sprayed on a nickel alloy wire or ribbon and for indirectly-heated cathodes on a metallic cylinder

The operating temperature is low, from 800 to 900°C. As a result the emission efficiency is high, ranging from 10 to 125mA per sq. cm. per watt for an indirectly-heated cathode and from 200 to 1000mA per sq. cm. per watt for a directly-heated cathode. However, anode voltages above 500V will cause these cathodes to be seriously damaged by positive-ion bombardment. Their use is thus restricted to low-power transmitting valves, and all types of receiving valves.

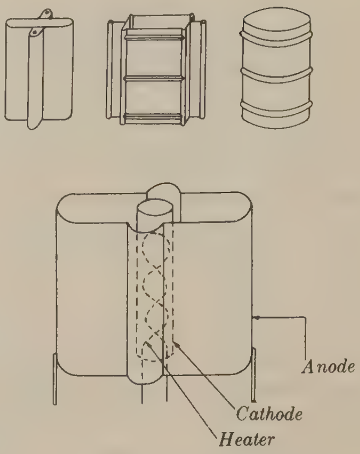


Fig.1-16 Shapes of anodes.

1.8 Physical construction of the anode. Since the anode or plate is the electrode which attracts and receives the electrons emitted by the cathode, it is always operated at a higher positive potential than the cathode. Anode voltages range from about 20V to about 10KV depending on the use and type of valve.

Anodes are usually constructed as hollow cylinders with various cross-sectional shapes. Some typical anode shapes are shown in Fig. 1-16. The cathode is placed along the axis of the anode cylinder to ensure that electrons are drawn uniformly from the entire cathode surface. In Fig. 1-16 an anode and cathode are shown together in their usual relative positions.

The size and material of the anode depend on the maximum power to be dissipated as heat at its surface. This heat is generated by the electron bombardment and by the passage of the electron current and must be safely radiated to prevent the anode temperature from rising to its melting point. For a given

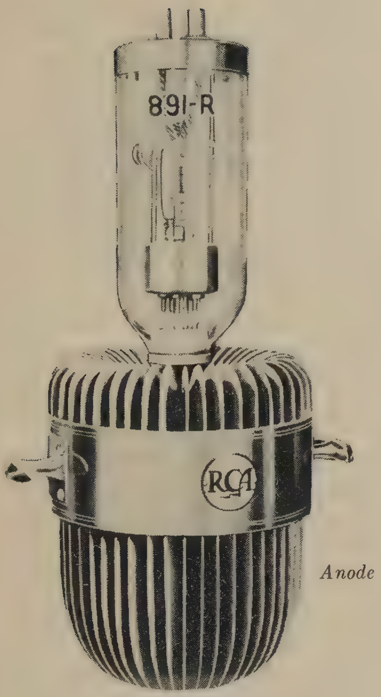


FIG.1-17 Air-cooled valve.

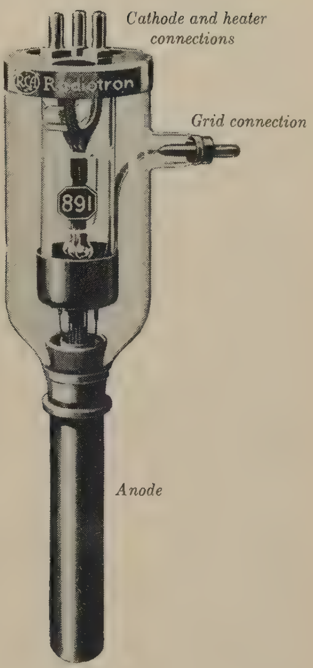


FIG. 1-18 Water-cooled valve.

material the amount of heat radiation depends on the area, color and structure of the surface. The dissipation may be increased by roughening and blackening the surface, since a rough dark surface radiates a greater amount of heat than a smooth shiny surface. Further, the power dissipation of the anode may be increased by increasing its surface area. To keep the anode from being unwieldy, its surface may be increased by attaching metal fins, across which a current of cooling air is blown. The fins are attached to a portion of the anode which is made to project outside the glass seal. The air-cooled valve shown in Fig. 1-17 has a *maximum anode dissipation* of $4000W$.

The power dissipation of an anode can also be increased by the continuous passage of a stream of cold water over its outer surface. The anode, usually of copper, extends through the glass seal into the cooling jacket. The water-cooled valve shown in Fig. 1-18 has a maximum anode dissipation of 6000 watts. The anode in this valve is a metal cylinder which extends below the glass and contains the other electrodes.

The anode material should possess the following properties.

(i) A high melting point to allow for high anode dissipation.
(ii) No tendency to warp and cause the electrode spacings and hence the electrical characteristics of the valve to change.
(iii) High vaporization temperature to prevent loss of anode material through evaporation and consequent metallic deposits on other electrodes and insulators.
(iv) Little tendency towards occlusion of gases.
(v) Malleability for ease in shaping.
(vi) Good thermal conductivity.

The following substances are most widely used in the construction of anodes:

1. Tungsten.

Tungsten was one of the first substances to be used for anodes. It has a high melting point, no tendency to warp and is very rugged. However it is extremely difficult to work into a desired shape.

2. Molybdenum.

This metal also has a high melting point. Since it is more malleable than tungsten, it is suitable for use in medium- and high-power transmitting valves. Since it is difficult to produce sheet molybdenum, anodes of molybdenum are usually basket-like in construction.

3. Carbon.

Carbon (graphite) conducts heat readily, is easily worked and does not warp at operating temperatures. It has a greater tendency towards occlusion of gases than other materials although careful selection and processing produce satisfactory results. For mechanical strength a carbon anode must be made with thick walls.

4. Tantalum.

This metal is similar to molybdenum. It has a higher melting point and greater malleability than molybdenum. Its chief advantage is that it has the ability to retain any occluded gases.

5. Nickel.

Nickel is inexpensive and malleable. However it tends to warp and has a low melting point. To improve its radiation, a layer of black, rough-surface amorphous carbon is deposited on its surface.

6. Copper.

Since copper is an excellent conductor of heat it is used for water-cooled and air-cooled anodes.

7. Zirconium-Coated Anode.

This type of coating gives high heat-dissipating characteristics to the valve. It also has the ability to retain occluded gases and absorb any other traces of gas in a valve. Therefore it maintains a high vacuum and a valve with such an anode can withstand heavy temporary overloads without damage to its filament from positive-ion bombardment.

1.9 The envelope and its evacuation. To maintain a high vacuum between the electrodes of a valve, they are enclosed in a glass or metal casing called an *envelope*. Metal envelopes are usually smaller than corresponding glass envelopes because they conduct heat out of the valves more readily. Glass envelopes are used in all high-power r-f valves because energy losses due to eddy currents would be excessive in metal envelopes. The use of metal envelopes and metal-sprayed glass envelopes is restricted to receiving valves and low-power transmitting valves. In glass or metal valves the leads from the electrodes pass through a glass seal at the bottom of the evacuated envelope to connecting pins and a composition base covers the glass seal. Typical base constructions are shown in Fig. 1-19 (a) and (b) respectively.

The composition base may be a ceramic or plastic material. Glass and porcelain are examples of ceramic bases, while bakelite is the commonly used plastic. Bakelite is inexpensive and is satisfactory for most receiving valves. When high r-f voltages exist between pins in the base, the ceramics are preferable because of lower dielectric loss.

The evacuation process for a valve consists of two steps:

(i) the exhausting of the envelope

(ii) the removal of occluded gases from the valve electrodes.

These processes are brought into operation after the inside electrodes and insulators have been assembled and the base and envelope have been sealed or welded together.

The envelope is initially exhausted by pumping the air out through the exhaust tube in the base of the envelope. At the

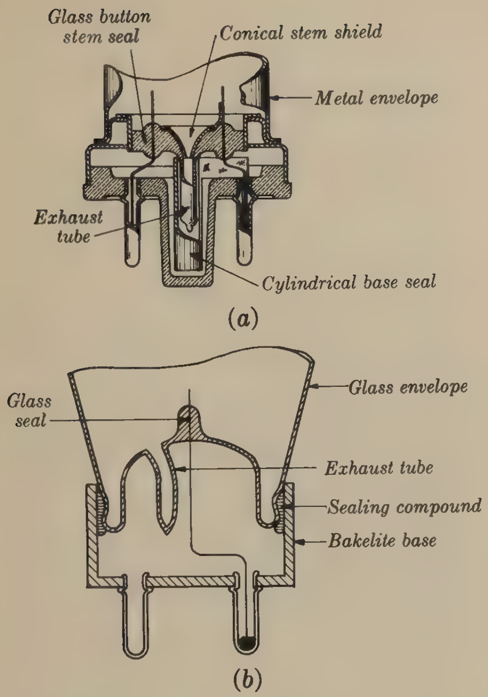


FIG. 1-19 (a) Metal envelope base construction (b) Glass envelope base construction.

same time the valve is heated to ensure that all gases are liberated as completely as possible.

Most electrode materials retain slight amounts of gas in their surface pores after the initial exhaustion process. At operating temperatures these occluded gases are expelled from the electrodes and the gas pressure in the valve is increased. Occluded gases are removed by either gettering or electrode heating.

In the *gettering process*, a small piece of an easily vaporized metal such as magnesium or barium is fastened to the inside structure of the valve near the base. This piece is called the *getter*. The valve is placed inside a solenoid through which high-frequency currents are passed. The eddy currents induced in the getter heat it and cause it to vaporize suddenly. The vapour condenses around the inside of the valve forming a silvery coating. The electrodes which are also heated by induction give up their occluded gases. The freshly condensed getter absorbs these gases making the vacuum more complete. In their new position the absorbed gases are not subject to as great heat as if they had remained occluded in the active elements of the valve. Some getters act as keepers since they will combine chemically with any occluded gases which may be liberated in normal operation. The gettering is done either immediately before or after sealing the exhaust tube.

In one *electrode-heating process*, the electrodes in a glass valve are heated by an induction coil. High-frequency currents are passed through a coil surrounding the valve. The eddy currents induced in all the metal electrodes heat them very rapidly. After a few seconds the sealing off may be completed. However in a metal valve the electrodes can not be heated by the induction method since the metal shell acts as a magnetic shield reducing the high-frequency field within the valve to a negligible amount.

The electrodes in large transmitting valves are heated by an electron-bombardment method. The tungsten filament is heated and high potentials are applied to the other electrodes in turn. Under the influence of the electric fields produced, the emitted electrons attain high velocities. Their kinetic energies are converted into heat as they strike the electrode. This method may be applied only after the initial exhausting has been completed, otherwise positive-ion bombardment may damage the cathode. The slight traces of gas remaining after the first process prevent this method from being used with oxide-coated valves.

Getters are not used in high-power valves since these valves become so hot in operation that the getter would vaporize again and destroy the vacuum. Such valves are evacuated by a combination of heating and exhaustion over long

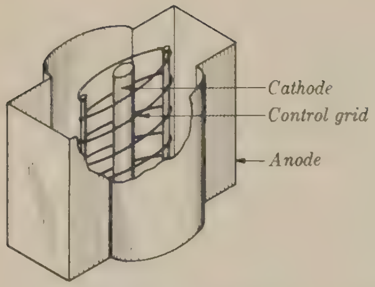


FIG.1-20 Arrangement of electrodes in a triode.

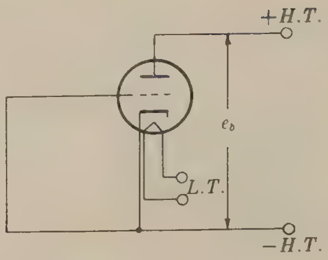


FIG.1-22 Triode arranged to operate as a diode.

periods. Since some gas remains, oxide-coated cathodes can not be used because of positive-ion bombardment. For water- or air-cooled valves in which copper anodes are used, exhaustion is less thorough for mechanical reasons and tungsten cathodes are required to withstand the positive-ion bombardment which occurs in operation.

In some valves, a certain amount of gas such as mercury vapour is required. The pressure in such valves is very low, approximately 10^{-8}mm. of mercury. Gas-filled valves are discussed in section 1.26.

1.10 The triode. A triode is a valve which contains a third electrode, called the *control grid*. This control grid takes the form of a cylindrical open helix or mesh of fine wire and is mounted to surround the cathode. In the construction of the triode the control grid is placed nearer to the cathode than to the anode (Fig. 1-20). If this grid is operated at zero potential it has little or no effect on the anode current of the valve since the electrons moving from cathode to anode pass freely through it due to the open nature of its construction. However, if the potential of this grid is made negative with respect to the cathode, the grid repels some electrons to the cathode and the anode current is reduced. Further, if the grid is made positive with respect to the cathode, an accelerating influence is exerted on the electron stream. Since the grid is nearer to the cathode than to the anode, the potential applied to the grid is much more effective in controlling the electron stream than is the anode potential. Thus this new electrode, the control grid, gets its name from its purpose and construction in the valve. The circuit symbol of a triode is shown in Fig. 1-21.

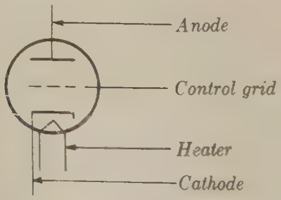


FIG.1-21 Circuit symbol for a triode with indirectly-heated cathode.

A more complete understanding of the operation of the triode and the function of the control grid may be obtained by a consideration of the circuits shown in Figs. 1-22, 1-23, 1-24. When the L.T. and H.T. supplies are connected as shown in Fig. 1-22, anode current flows in the anode circuit of the valve. Since the grid is connected to the cathode, the grid and cathode are at the same potential and the effect of the grid on the electron stream is negligible. As a result the valve acts as a diode. If, however, a grid voltage supply (E_{cc}) is now connected across grid and cathode via the voltage divider R, as shown in Fig. 1-23, the grid is made negative to the cathode. When the voltage divider tapping point is set to make the grid slightly negative with respect to the cathode it repels some

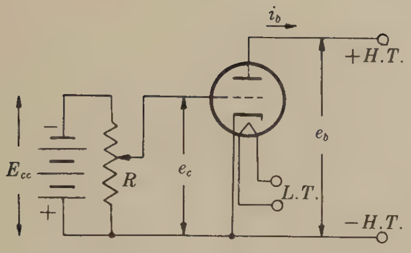


FIG.1-23 Triode circuit arranged to make control grid negative with respect to the cathode.

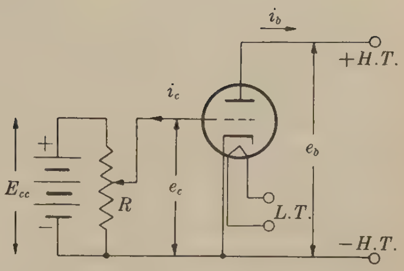


FIG.1-24 Triode circuit arranged to make control grid positive with respect to the cathode.

of the electrons and the anode current (i_b) is reduced. Since the grid is between the cathode and anode, the grid voltage (e_c) has a greater effect on the anode current (i_b) than has the anode voltage (e_b). Thus a small negative grid voltage (e_c) is capable of decreasing the anode current considerably. If the grid is made more negative by variation of R, more electrons are repelled to the cathode and the anode current is further decreased. Finally, if the control grid is made sufficiently negative all electrons emitted by the cathode are repelled and none reach the anode. Under this condition the anode current is cut off. The exact negative grid voltage required to reduce the anode current to zero is called the *cut-off* voltage.

If the grid is connected across the supply as shown in Fig. 1-24 the grid is made positive with respect to the cathode. When the grid voltage is positive the electrons are attracted and they receive extra acceleration on their way to the anode. As a result of this acceleration most of the electrons pass through the grid. Thus the grid also controls the anode current when it is positive. However, some electrons attracted by the grid are collected by it and return through the external circuit of the grid supply to the cathode. This flow of electrons is known as the *grid current* of the valve and the path it takes is referred to as the *grid circuit.* The value of grid current flowing at any instant is represented by the abbreviation i_c. It is so small that a microammeter would be required to measure it.

Should the anode voltage be high enough to produce saturation with zero grid voltage, placing a positive potential on the grid cannot increase the anode current. Instead, grid current will flow and as extra electrons are not available from the cathode the anode current will decrease by an amount equal to the grid current.

1.11 The characteristic curves of a triode. The inclusion of the control grid in the triode valve makes it possible to control the flow of current from cathode to anode. A small variation in voltage applied to the grid results in a comparatively large variation in current flowing through the valve. If the current is made to flow through a resistor connected to the anode, corresponding large variations in voltage are developed across the resistor. Hence small changes in input voltage applied to the grid produce large changes in output voltage from the anode. In this way the valve circuit acts as a voltage amplifier.

A continuous picture of the relationship between the grid voltage, anode current and anode voltage of a triode may be obtained by plotting the *characteristic curves* of the valve. Since there are three variables it is necessary to investigate the following relationships.

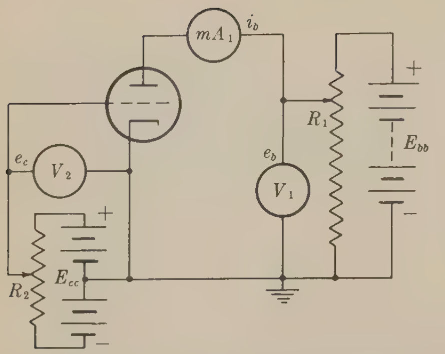


FIG.1-25 Circuit for obtaining data to plot the characteristic curves of a triode.

(i) The relationship between the anode voltage (e_b) and the anode current (i_b) with the grid voltage constant (E_c). The graphical representation of this relationship is called the $e_b - i_b$ curve or the *static anode characteristic* of the triode as in the case of the diode.

(ii) The relationship between the grid voltage (e_c) and the anode current (i_b) with the anode voltage constant (E_b). The graphical representation of this relationship is called the $e_c - i_b$ curve or the *static mutual characteristic* of the triode.

Each of these relationships may be investigated by means of the circuit shown in Fig. 1-25. Potentiometer R_1 permits the anode voltage (e_b) to be varied from zero to the maximum voltage, E_{bb}, of the H.T. supply. Voltmeter V_1 measures the anode voltage and milliammeter mA_1 measures the anode current. Potentiometer R_2 permits the grid potential to be varied from a negative value to a positive value with respect to the cathode and voltmeter V_2 measures the grid voltage.

A. The Static Anode Characteristics or $e_b - i_b$ Curves

To investigate the relationship between the anode voltage and anode current the grid voltage is first set at 0 volts and then as the anode voltage is varied in steps from 0 volts up to full anode voltage rating of the valve, the corresponding values of anode current are recorded. The values of anode voltage are used as abscissae and the values of anode current as ordinates to plot points. The characteristic is obtained by drawing a smooth curve through the plotted points (Fig. 1-26). A family of anode characteristic curves as shown in Fig. 1-26 may be obtained by repeating the above procedure for various values of E_c.

The individual $e_b - i_b$ curves of the family have the same general shape and are almost parallel. The general shape of each static $e_b - i_b$ curve of a triode is similar to that of a diode. As the grid potential is made more negative, more electrons are repelled to the cathode; hence the values of anode current are decreased for the same values of anode voltage. Thus, increasing the negative grid voltage results in a curve farther to the right in the diagram but the general shape remains much the same. It should be noted that the upper portion of each curve is almost a straight line. Hence for this portion of each characteristic there is a linear relationship between anode current and anode voltage.

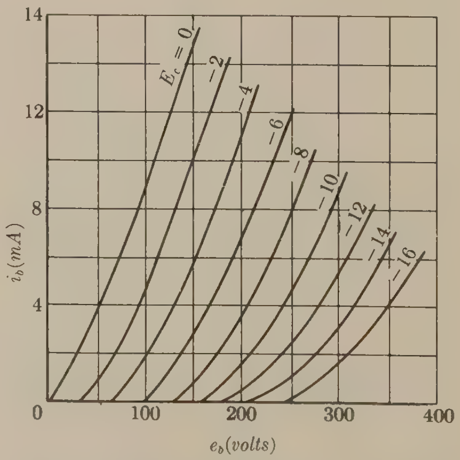


FIG.1-26 Triode static anode characteristics ($e_b - i_b$ curves).

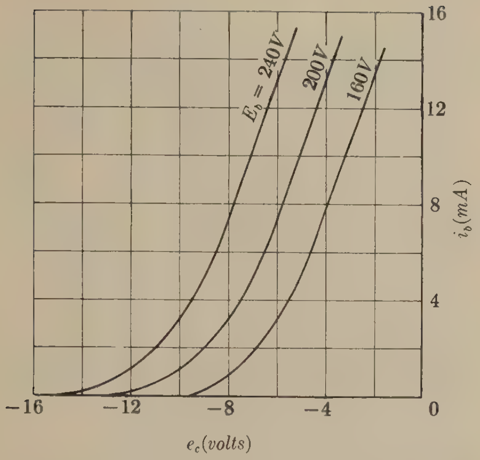


FIG. 1-27 Triode static mutual characteristics ($e_c - i_b$ curves).

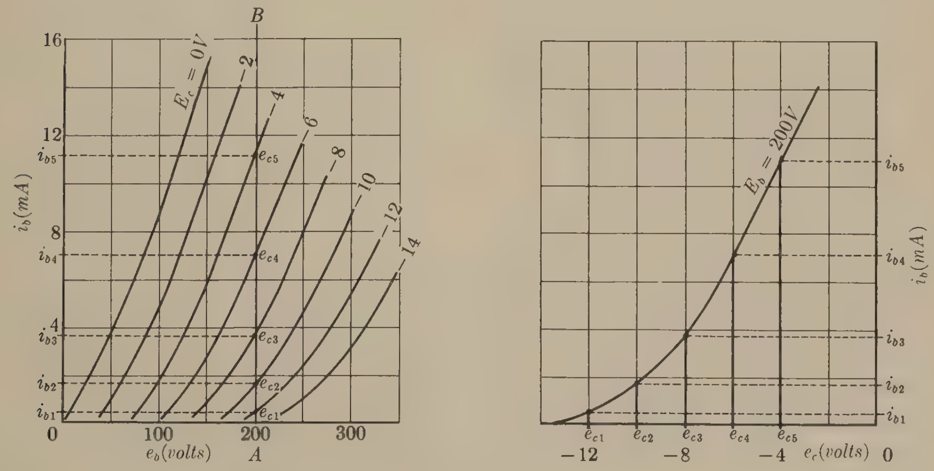
B. The Static Mutual Characteristics or $e_c - i_b$ Curves

To investigate the relationship between the grid voltage and anode current, the anode voltage (Fig. 1-25) is maintained constant at a suitable value by means of the potentiometer R_1. The grid voltage indicated on V_2 is varied in steps from beyond cut-off to about zero by means of potentiometer R_2. The corresponding values of anode current are read from mA_1 and recorded. The values of grid voltage are used as abscissae and the values of anode current as ordinates to plot the mutual characteristic curve (Fig. 1-27). Other curves of the family are obtained by repeating the above procedure with the anode voltage held constant at various other values (Fig. 1-27).

The shapes of all $e_c - i_b$ curves of the family in Fig. 1-27 are similar. However as the constant anode voltage is reduced, fewer electrons are attracted to the anode and hence the values of anode current are smaller for corresponding values of grid voltage. Thus the curve produced appears to the right as the constant anode voltage is reduced.

It should also be noted that the central portions of these curves are almost straight lines. Hence, over this portion of the static $e_c - i_b$ curve there is a linear relationship between the anode current and grid voltage.

Usually, only the static $e_b - i_b$ curves are given in valve manuals. However, when the static $e_c - i_b$ curves are desired they can be derived from the static $e_b - i_b$ curves as follows. Fig. 1-28 (a) shows a set of static $e_b - i_b$ curves for a triode. For a fixed anode voltage, E_b (200 V),



(a) (b)

FIG. 1-28 Method of deriving a static $e_c - i_b$ curve from the static $e_b - i_b$ curves.

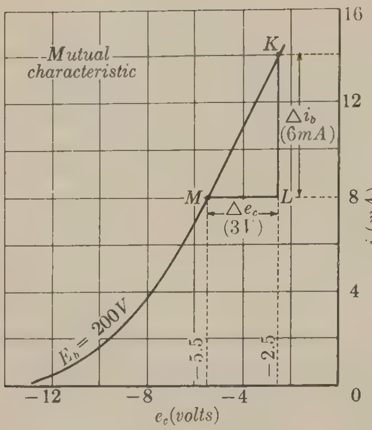


Fig. 1-29 Method of obtaining mutual conductance of a triode.

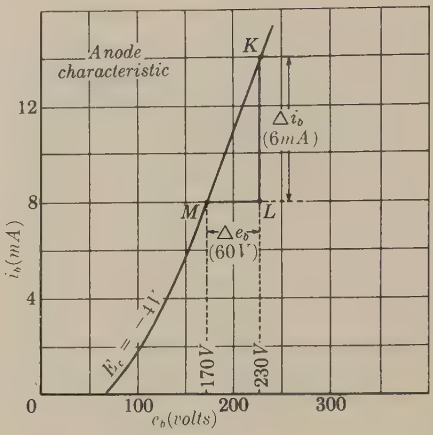


Fig. 1-30 Method of obtaining a-c anode resistance of a triode.

the values of anode current corresponding to several grid voltages can be obtained by drawing a vertical straight line AB to cut across the given curves. The points of intersection give corresponding values of anode current, i_{b1}, i_{b2}, etc., and grid voltage, e_{c1}, e_{c2}, etc. These i_b and e_c values are plotted to give a static $e_c - i_b$ curve as in Fig. 1-28 (b). This process, when repeated, gives static $e_c - i_b$ curves for other fixed values of anode voltage and a family of static $e_c - i_b$ curves can be obtained.

1.12 Valve factors. Three important factors or ratios which may be obtained from either set of characteristic curves, are useful when analyzing the performance of a valve. These valve factors are:

(i) *the amplification factor*
(ii) *the mutual conductance*
(iii) *the a-c anode resistance.*

Each of these factors depends upon the construction of the valve and upon the electrode voltages and currents and helps to indicate the ability of a given valve to perform a specific function. Although these factors are discussed as applied to a triode, the general definitions apply equally well to valves with more than three electrodes.

A. *The Amplification Factor*

The *amplification factor* of a valve is a measure of the effectiveness of the grid as compared to the anode in controlling the flow of anode current. Thus, in general, the amplification factor, designated by the Greek letter μ, is the ratio of the change in anode voltage required to cause a small change in anode current to the change in grid voltage required to cause the same small change in anode current.

Thus $\mu = \dfrac{\triangle e_b}{\triangle e_c}$ for the same small change in i_b

where μ = amplification factor
$\triangle e_b$ = a small change in anode voltage which produces a small change in anode current with the grid voltage constant
$\triangle e_c$ = a small change in grid voltage which produces the same small change in anode current with the anode voltage constant.

The amplification factor of the triode, the characteristics of which are shown in Figs. 1-29 and 1-30, is determined as follows.

Right-angled triangles KLM are drawn with the linear portion of each curve as hypotenuse. By means of each of these triangles KLM the same change in anode current from $8mA$ to $14mA$ is measured on each curve. Thus $\triangle i_b$ is $6mA$. The grid voltage change $\triangle e_c$ is $3V$ and the anode voltage change $\triangle e_b$ is $60V$.

$$\text{Then} \quad \mu = \frac{\triangle e_b}{\triangle e_c} = \frac{60}{3} = 20.$$

Therefore the amplification factor is 20 in the range of $8-14mA$ anode current and an anode voltage of $200V$.

In this range the grid voltage is 20 times more effective in controlling the anode current than is the anode voltage. It is this property of a control grid which makes amplification possible (Chapter II). For triodes the value of the amplification factor ranges from about 4 to about 100.

The amplification factor obtained above was determined only for the linear portion of the characteristics (Fig. 1-29 and Fig. 1-30). This is the most important range, since a triode is usually operated with values of anode and grid voltage which lie on the straight portion of the characteristic. In this region the valve factors do not change appreciably. As a result, the values of the valve factors which are given in the makers' specifications are those which are obtained from the linear range and hence only hold true for this region of operation. For this reason the valve factors are sometimes referred to as *valve constants*. When desired, the valve factors may be calculated in a similar way when the operating voltages lie on the curved portions of the characteristic curves.

B. The Mutual Conductance

The *mutual conductance* or *grid-to-anode transconductance*, symbolized by g_m, is a measure of the effectiveness of the grid in controlling the anode current. It is defined as the ratio of a change in anode current to the change in grid voltage required to cause this change in anode current, for a constant anode voltage.

$$\text{Thus,} \quad g_m = \frac{\triangle i_b}{\triangle e_c} \quad \text{if } E_b \text{ is constant,}$$

where g_m = transconductance in mhos
$\triangle e_c$ = a small change in grid voltage in volts
$\triangle i_b$ = a small change in anode current, resulting from $\triangle e_c$, in amperes.

The mutual conductance can be calculated for the linear portion of the static $e_c - i_b$ curve of Fig. 1-29 as follows: From triangle KLM,

$$\triangle i_b = LK = 6mA$$

$$\triangle e_c = LM = 3V, \text{ for } E_b = 200V$$

$$\text{Thus} \quad g_m = \frac{\triangle i_b}{\triangle e_c} = \frac{LK}{LM} = \frac{6\times10^{-3}}{3} = 2\times10^{-3}$$

Therefore, the transconductance is 2×10^{-3} mhos, or 2000 micromhos. Further, if $\triangle e_c$ is 1 volt,

$$\text{since } \frac{\triangle i_b}{\triangle e_c} = .002,$$

$$\text{then } \triangle i_b = .002A \text{ or } 2mA.$$

Therefore, a 1-volt change in the grid voltage causes a $2mA$ change in the anode current, if the anode voltage is constant. As the value of the transconductance of the valve increases a 1-volt change in the grid voltage causes a still larger change in the anode current. Thus a large transconductance means that the grid voltage has a great amount of control over the anode current. From Fig. 1-29 it may be seen that the value of g_m is equal to the slope of the linear portion of the curve, since $\frac{LK}{LM}$ is the slope. Hence an $e_c - i_b$ curve with a steep slope indicates a large transconductance. Values for the transconductance of a triode range from about 1000 to about 5000 micromhos.

Since the slopes of all curves of the mutual characteristic family in Fig. 1-27 are for all practical purposes the same, that is, the straight portions are parallel, the value of the transconductance is the same whether the anode voltage is constant at 240, 200 or 160V.

C. The A-C Anode Resistance

As in the case of a diode, the *a-c anode resistance* (r_a) of a valve is the resistance that it offers to the flow of a small alternating current which is superimposed on some steady anode current. Thus, the a-c anode resistance is a measure of the effectiveness of the anode voltage in controlling the anode current. In general, it is defined as the ratio of a small change in the anode voltage to the change in anode current which it produces, for a constant grid voltage.

Thus, $r_a = \dfrac{\Delta e_b}{\Delta i_b}$ if E_c is constant,

where r_a = anode resistance in ohms

Δe_b = a small change in anode voltage in volts

Δi_b = a small change in anode current, caused by Δe_b, in amperes.

The anode resistance is measured in ohms, since it is a ratio of voltage to current. When a valve is operating on the linear portion of its characteristic the anode resistance may be determined from the values shown in Fig. 1-30. The values of Δe_b and Δi_b are obtained as before by drawing the triangle KLM. Thus from the graph $\Delta e_b = LM = 60V$ and $\Delta i_b = LK = 6\ mA$ for $E_c = -4$V.

$$\text{Then} \qquad r_a = \frac{\Delta e_b}{\Delta i_b} = \frac{LM}{LK} = \frac{60}{6\times10^{-3}} = 10{,}000.$$

Therefore the anode resistance is 10,000 ohms. Further, if Δe_b is 1V,

$$\text{since} \qquad \Delta i_b = \frac{\Delta e_b}{r_a}$$

$$\text{then} \quad \Delta i_b = \frac{1}{10{,}000}\ A = 0.1mA.$$

Thus a change of 1 volt in the anode voltage causes a change of 0.1mA in the anode current. A large anode resistance therefore means that the change in anode current is very small for a change in anode voltage or in other words that the anode voltage exerts little control over the anode current.

Since the linear portions of all curves of the family of anode characteristics of Fig. 1-26 are almost parallel, the value of anode resistance is practically the same for the other values of E_c shown. Since the slope of the curve in Fig. 1-30 is $\dfrac{LK}{LM}$ and $r_a = \dfrac{LM}{LK}$, therefore r_a is the reciprocal of the slope of the $e_b - i_b$ curve. Thus a characteristic with a steep slope has a low anode resistance.

For triodes, the values of anode resistance range from about 1000 to about 25,000 ohms.

D. Calculation of the Valve Factors for Operation on Curved Portions of the Characteristics

The method of calculation of valve factors for the curved portions is very similar to that used for the straight portions. It has been pointed out that the transconductance is equal to the slope of the static $e_c - i_b$ curve and that the anode resistance is equal to the reciprocal of the slope of the static $e_b - i_b$ curve. The values of the valve factors may then be calculated at points along the curved portions by measuring the slopes at corresponding anode current points. The necessary steps in obtaining the slopes of the curves and hence the valve factors are shown in Fig. 1-31. The curves shown in this diagram are members of the same family of characteristics as the curves of Figs. 1-29 and 1-30. The anode current point P chosen on each characteristic is 4mA. This lies on the curved portions of the $e_c - i_b$ and $e_b - i_b$ curves shown in Fig. 1-31 (a) and (b) respectively. The slope of each curve at P is the slope of

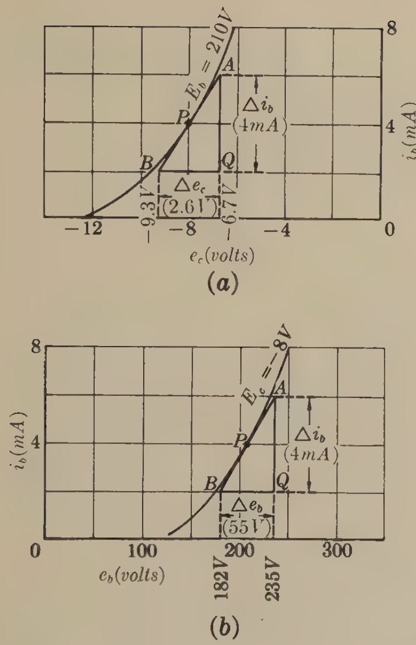


FIG. 1-31 Method of calculation of values of valve factors for operation on curved portions of the characteristics.

the tangent AB drawn at this point. A right-angled triangle QAB is constructed as in the previous calculations. From triangle QAB in Fig. 1-31 (a) the slope at point P is

$$\frac{AQ}{BQ} = \frac{4\times10^{-3}}{2.6} = 1.54\times10^{-3}$$

Therefore the transconductance is 1.54×10^{-3} mhos or 1540 micromhos. This value is correct only if the grid voltage is $-8V$ and the anode voltage is $210V$.

The anode resistance is found from triangle QAB in a similar manner in Fig. 1-31 (b) for a grid voltage of $-8V$ and an anode voltage of $210V$ (point P on the curve). The slope of the curve at the point P is $\frac{AQ}{BQ}$ and the reciprocal of the slope is $\frac{BQ}{AQ}$. Hence the anode resistance is given by

$$\frac{BQ}{AQ} = \frac{55}{4\times10^{-3}}$$

$$\text{or}\quad \frac{BQ}{AQ} = 13.7\times10^{-3}$$

$$= 13{,}700$$

Therefore the anode resistance is 13,700Ω.

The values obtained above for g_m and r_a hold true only at the point for which they are calculated. It is seen that g_m changed from 3000 micromhos to 1540 micromhos and r_a changed from 10,000 to 13,700 ohms from the straight portion to this point on the curved portion of the characteristics.

E. The Relationship Between the Valve Factors

The relationship between the valve factors is $r_a\times g_m = \mu$. This is proved as follows.

By definition,

$$g_m = \frac{\Delta i_b}{\Delta e_c} \text{ with } E_b \text{ constant,}$$

$$r_a = \frac{\Delta e_b}{\Delta i_b} \text{ with } E_c \text{ constant,}$$

$$\mu = \frac{\Delta e_b}{\Delta e_c} \text{ for the same value of } \Delta I_b.$$

Then

$$r_a\times g_m = \frac{\Delta i_b}{\Delta e_c}\times\frac{\Delta e_b}{\Delta i_b}$$

$$= \frac{\Delta e_b}{\Delta e_c}$$

$$= \mu$$

Thus $\mu = r_a\times g_m$

where μ = amplification factor

r_a = anode resistance in ohms

g_m = transconductance in mhos.

The amplification factor of the triode over the straight portions of the characteristic curves shown in Figs. 1-29 and 1-30 can be calculated from this formula. Since r_a = 10,000 ohms and g_m = .002 mhos, then

$$\mu = r_a\times g_m$$

$$= 10{,}000\times.002$$

$$= 20$$

Therefore the amplification factor is 20. This is the same value as previously calculated by another method.

The amplification factor for the operating point P, Fig. 1-31 (a) and (b), is calculated in a similar manner.

Since $r_a = 13{,}700$ ohms and $g_m = 1540$ micromhos for $E_c = -8V$ and $E_b = 210V$,

$$\text{then} \quad \mu = r_a \times g_m = 13{,}700 \times 1540 \times 10^{-6} \doteqdot 21.$$

Therefore the amplification factor is approximately 21; this value applies only for $E_c = -8V$ and $E_b = 210V$.

1.13 Amplification. In Chapter I, Vol. 1, the energy transformations involved in the various radio systems are discussed. Somewhere in each of these systems it becomes necessary to amplify small a-c voltages and currents. For example, in a radio transmission system, sound energy is first converted into electrical energy by means of a microphone. This electrical energy appears in the form of a very low-power alternating current and hence it becomes necessary to amplify this a-c signal before it is changed into energy in the form of radio waves. By amplification the radiated energy can be made large enough to make communication possible over great distances. However, since energy cannot be created it is not possible to increase the electrical energy of the system without taking energy from some other source. In an electrical amplifier the a-c signal is used to control the energy output of some other source such as an H.T. supply. A valve having a control grid provides a means for controlling the energy from such a source and so makes amplification of a-c voltages and currents possible.

The characteristic curves of a valve provide the necessary data from which its performance in an amplifier circuit can be determined. In amplifier circuits the a-c signal voltage is applied between grid and cathode of the valve and so it is sometimes referred to as the *grid-input* or *grid-signal* voltage while the *grid circuit* is referred to as the *input circuit.*

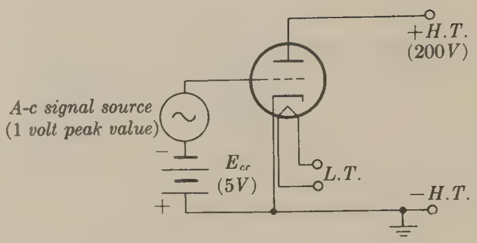


FIG. 1-32 Triode with a-c input voltage.

The operation of a valve with an alternating input voltage may be understood by a consideration of the circuit of Fig. 1-32. The anode is connected to the positive terminal (+H.T.) of a high-voltage d-c source. The cathode is connected to the negative terminal (−H.T.) which is at ground potential. The grid is connected to the negative terminal of a low-voltage (5V) battery through a source of alternating input voltage. The cathode is connected to the positive terminal of this battery.

When the grid-signal voltage is zero, the grid is held at a steady negative potential of $-5V$ relative to the cathode. From the $e_c - i_b$ curve of Fig. 1-33 it can be seen that with an initial grid voltage of $-5V$ the anode current is $10mA$ and the valve is operating at point A on the curve.

When the grid-signal voltage of 1-volt peak value is applied in series with the −5 volts as shown, the voltage between grid and cathode swings between −4 and −6 volts, alternate half cycles subtracting from and adding to the initial (negative) grid voltage. At the same time, the anode current increases and decreases in accordance with the changes in grid voltage. Thus it varies between the limits $13mA$ and $7mA$ and from Fig. 1-33 it may be seen that the waveform of the anode current variation is an exact reproduction of the waveform of the a-c grid-signal voltage. Thus, a 1-volt peak a-c signal causes an anode current variation of $6mA$. This varying

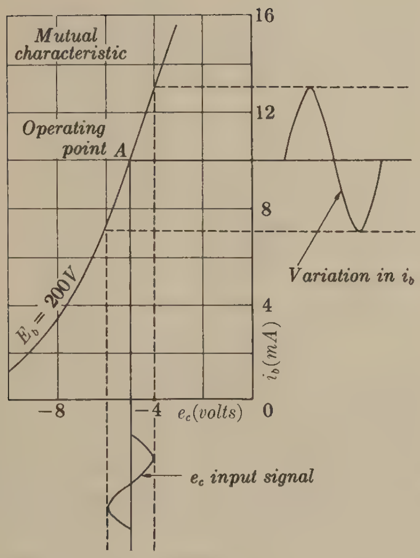


FIG. 1-33 Input and output waveforms of triode circuit shown in Fig. 1-32.

current is equivalent to the original direct current with an alternating current superimposed on it. Thus the current is composed of a d-c average component, together with an a-c or varying component of peak value $3mA$.

The development of an alternating anode current in response to an a-c signal voltage is not enough to effect amplification of this voltage. In order to make use of the large variations in the anode current which result from the small variations in grid voltage, some *load* must be inserted in the anode circuit so that the alternating current through it can develop a relatively large or amplified voltage across it.

Fig. 1-34 shows the circuit of Fig. 1-32 with a resistor R inserted as an *anode load.* The anode current variation produces a voltage variation across the anode load resistor. This voltage variation is called the *output voltage* of the circuit. By means of this circuit a small grid-signal voltage change produces a large output voltage. This process is the basis for all vacuum tube amplification and, although in principle it is simple, at the same time certain complications are introduced. Since the output voltage is developed across the load resistor R it is in effect developed at the anode of the valve and as a result with resistor R in the circuit it can no longer be assumed, as in the consideration of the circuit of Fig. 1-32 (a), that the anode voltage is constant. Instead it increases and decreases with the alternations of the grid voltage. Thus the static characteristics do not give complete information to indicate how a valve operates in a practical circuit. However, the operation of a valve with a resistance load can be represented graphically but the curve depends not only on the H.T. voltage and the tube characteristics but also on the load resistance. Such a curve is known as the *dynamic characteristic* of the valve and the method of constructing it is considered in Chapter II which discusses the topic of amplification in detail.

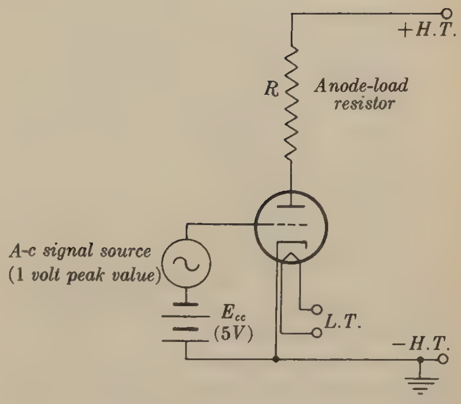


FIG. 1-34 Simple triode amplifier circuit.

1.14 Interelectrode capacitance. Capacitance exists between any two pieces of metal which are separated by an insulating dielectric. Thus there is a certain amount of capacitance present between any two electrodes of a vacuum tube. The capacitances that exist in a triode (Fig. 1-35) are

(i) the grid-to-cathode capacitance, C_{gk},
(ii) the grid-to-anode capacitance, C_{ga},

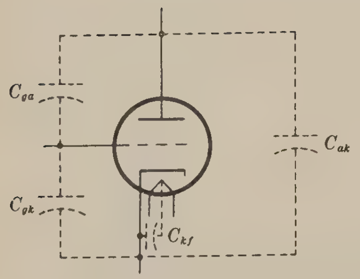


FIG. 1-35 Triode interelectrode capacitances.

(iii) the anode-to-cathode capacitance, C_{ak}
(iv) the cathode-to-filament capacitance, C_{kf}.

The values of these capacitances depend on the spacing and area of overlap of the electrodes. Typical values for a triode are:

$$C_{ga} = 2.0\ \mu\mu F$$
$$C_{gk} = 3.0\ \mu\mu F$$
$$C_{ak} = 11.0\ \mu\mu F$$

The high value for the anode-to-cathode capacitance, C_{ak}, is the result of the large area of overlap of cathode and anode. C_{ga} is relatively small because the grid is a mesh and so has little surface area. C_{gk} is larger than C_{ga} since the grid is much closer to the cathode than to the anode. The effective capacitances of a valve, measured when the electrodes are connected in a circuit, are greater than when the electrodes are disconnected. The increased interelectrode capacitance when the valve is in operation results from the shunting effects of the circuit wiring, tube base and socket.

1.15 Effects of interelectrode capacitance. Although the values of interelectrode capacitance are very small, they often introduce a coupling effect between input and output circuits associated with a valve. This makes the circuit unstable in its operation. The grid-to-anode capacitance introduces the greatest amount of unwanted coupling and its effect may be understood by a consideration of Fig. 1-36 which shows a simple triode amplifier circuit. In this circuit the grid-to-anode capacitance, C_{ga}, acts as a direct a-c coupling between the input circuit and the output circuit in parallel with the valve. As the value of C_{ga} is small, this shunting effect is negligible at low frequencies. However, as the frequency is increased, the shunting effect increases and becomes appreciable at radio frequencies. As a result of this effect the amplification is reduced at high frequencies. In addition the voltage developed across the output load resistor R is also directly coupled back into the input circuit by means of C_{ga}. As a result of this coupling energy is fed back from the output circuit to the input circuit. Under certain load conditions such *reaction* or *feedback* causes the amplifier to become unstable. If this unwanted coupling between input and output circuits is present to an unacceptable degree when a triode is used another type of valve is used in which a screen is present between grid and anode to reduce C_{ga}.

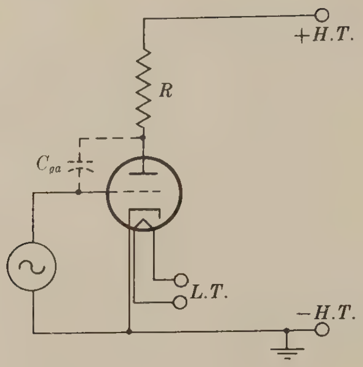


FIG. 1-36 Simple triode amplifier circuit.

1.16 The principle of screening. The capacitance of C_{ga} (Fig. 1-36) can be reduced by inserting an earthed metallic screen of sufficient size between its plates. Such a screen, consisting of a metallic plate larger than either of the two plates of C_{ga}, is shown in Fig. 1-37 connected through impedance Z to earth. The capacitance between g and a is replaced by two capacitances gs and sa in

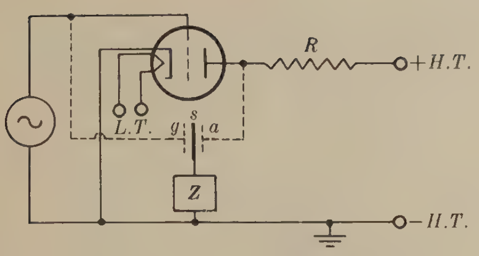


FIG. 1-37 Principle of screening.

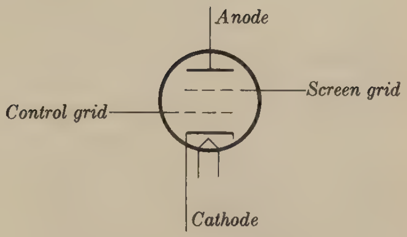


FIG. 1-39 Circuit symbol for a tetrode with an indirectly-heated cathode.

series. The current in the output circuit containing Z and R produces a p.d. across Z which is also in the input circuit and as a result current is still coupled back into the input circuit. If the impedance of Z is increased or made infinitely large by disconnecting it from the circuit, the coupling is further increased. However, if s is connected directly to earth, current flows as before in the output circuit, but since there is no impedance common to both output and input circuits there is no voltage and therefore no current developed in the input circuit. Hence, no matter what current flows in the output circuit and no matter what voltages appear across sa no current is fed back into the input. As a result g is completely screened from a. The screen as is only effective if it entirely isolates g from a and perfect screening results only when s is connected to earth either directly or through an impedance which is negligibly small.

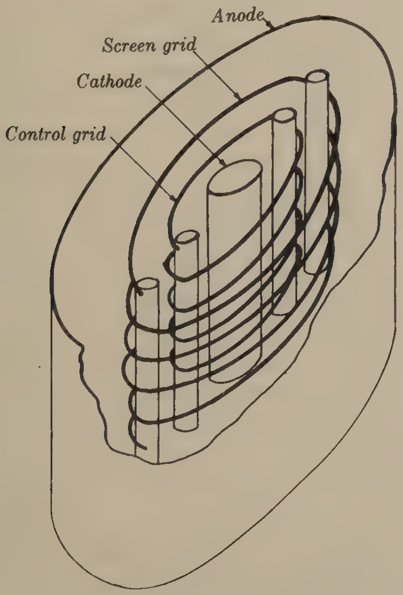


FIG. 1-38 Arrangement of electrodes in a tetrode.

The principle of screening is used in *tetrodes* to reduce appreciably the grid-to-anode capacitance encountered in triodes. A wire mesh similar to the control grid in shape and construction is mounted so that it completely surrounds the control grid (Fig. 1-38). It is evident that this mesh will not provide complete screening as would solid metal. However, a solid screen would cut off the flow of electrons from cathode to anode.

The new electrode is called a *screen grid*. A valve containing the four electrodes, screen grid, anode, cathode and control grid is called a *tetrode* or sometimes a *screen-grid valve*.

Where C_{ga} in a triode is of the order of $2 \mu\mu F$, in tetrodes it is of the order of $.002\ \mu\mu F$.

1.17 The tetrode. The *tetrode* with its four electrodes, cathode, control grid, *screen grid* and anode is represented by the circuit symbol shown in Fig. 1-39.

If the screen grid of the tetrode is earthed as shown in Fig. 1-40, the potential of the screen grid is the same as that of the cathode. Hence electrons in the vicinity of the input side of the screen would not be drawn on to the positive anode and the anode current would fall almost to zero. However, the requirements of screening are satisfied if the

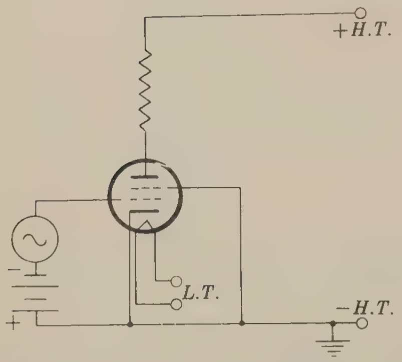


FIG. 1-40 Tetrode circuit with screen grid earthed.

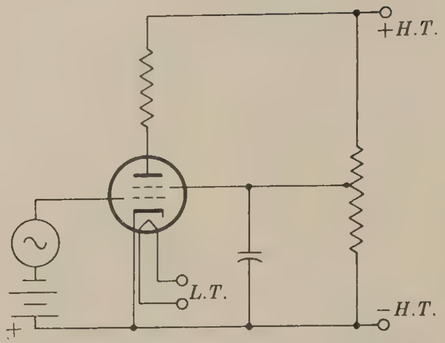


FIG. 1-42 Typical tetrode circuit arrangement.

screen is connected through a very low impedance to earth. Thus a capacitor of large capacitance connected from screen to earth (Fig. 1-41) effectively earths the screen for a-c. This permits the screen to be operated at a positive d-c potential by means of a connection to a voltage divider across the H.T. supply (Fig. 1-42). With the screen grid at a positive potential it collects some electrons. These flow from cathode to screen grid through the voltage divider and back to the cathode. This flow of electrons is called the *screen current* (i_{c2}) and the circuit is called the *screen circuit.*

1.18 The characteristic curves of a tetrode. A continuous picture of the relationship between the grid voltage, anode voltage, anode current and screen current of a tetrode may be obtained by plotting the characteristic curves of a tetrode. To show the relationship between these variables graphically, the following two characteristic curves are required.

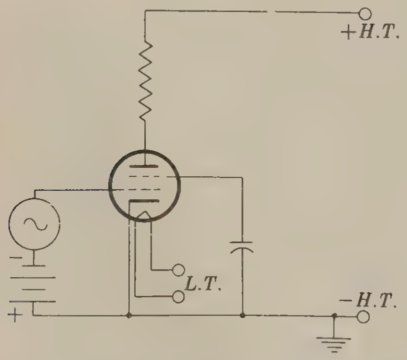


FIG. 1-41 Tetrode circuit with screen grid connected to earth via a capacitor.

(i) The static e_b-i_b curve or the static anode characteristic which presents the relationship between anode voltage (e_b) and the anode current (i_b) with the grid voltage (E_c) and screen voltage (E_{c2}) constant.

(ii) The static e_c-i_b curve or the static mutual characteristic which presents the relationship between grid voltage (e_c) and the anode current (i_b) with the anode voltage (E_b) and screen voltage (E_{c2}) constant.

Each of these relationships can be investigated by means of the circuit shown in Fig. 1-43. Potentiometer R_1 permits the anode voltage (e_b) to be varied from zero to the maximum voltage of the anode supply. Voltmeter V_1 measures the anode voltage and the milliammeter mA_1 measures the anode current. Potentiometer R_2 permits the grid potential to be varied from a negative voltage to a positive voltage with respect to the cathode. Voltmeter V_2 measures the grid voltage. Potentiometer R_3 permits the screen grid potential to be varied from zero to the maximum voltage of the screen supply. The milliammeter mA_2 measures the screen current.

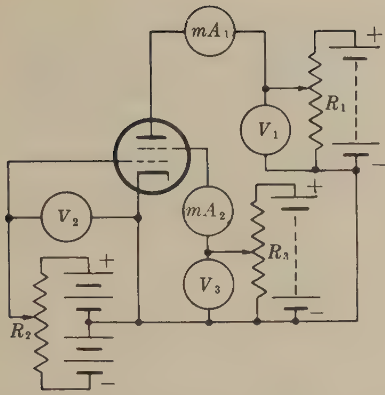


FIG.1-43 Circuit for obtaining data to plot the characteristic curves of a tetrode.

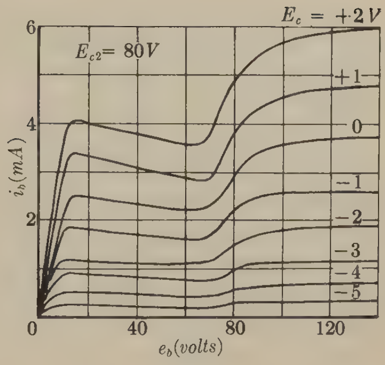


FIG.1-44 Tetrode static anode characteristics ($e_b - i_b$ curves).

A. *The Static Anode Characteristics or $e_b - i_b$ Curves*

To investigate the $e_b - i_b$ relationship the grid voltage (Fig. 1-43) is first set at a small positive value and screen grid voltage at its rated value. Then as the anode voltage is varied in steps from 0 volts up to the anode voltage rating of the valve, the corresponding values of anode current are recorded. The values of anode voltage are used as abscissae and the values of anode current as ordinates to plot points and so obtain the characteristic curve. By leaving the screen-grid voltage constant at the rated value, a family of anode characteristic curves (Fig. 1-44) may be obtained by repeating the above procedure for various values of E_c.

The individual $e_b - i_b$ curves of the family (Fig. 1-44) have the same general shape and are almost parallel. However, these curves are very different in shape from those of a triode (Fig. 1-26) indicating that the operation of this valve is quite different from that of a triode. An understanding of the operation of a tetrode may be obtained from a detailed study of one of these $e_b - i_b$ curves. The interpretation of an $e_b - i_b$ curve is made easier if, in the above procedure, the screen current is also recorded and these values plotted to obtain the corresponding $e_b - i_{c2}$ curve or *screen characteristic curve* as shown in Fig. 1-45. An understanding of the reasons for the variations in current shown by the curves of Fig. 1-45 provides in turn an understanding of the operation of a tetrode.

Fig. 1-45 shows the variation of anode current and screen current for a valve with screen-grid voltage maintained constant. When the anode voltage is zero, screen current is high, few electrons pass the screen grid, and anode current is insufficient to show on the graph with the scale used. As anode voltage is applied, more electrons are attracted through the screen to the anode and the anode current increases, at the expense of the screen current. The dip in the anode current curve is caused by secon-

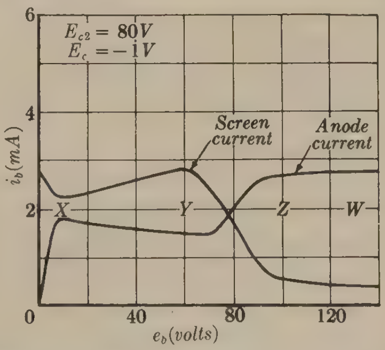


FIG.1-45 Tetrode static screen characteristic compared with anode characteristic.

dary electrons from the anode finding their way to the screen grid. When high-velocity electrons strike the anode they can dislodge loosely held electrons from the anode material. These latter are known as *secondary electrons*, and the process by which they are released is called *secondary emission*. Some of these electrons are dislodged with sufficient velocity to reach the region where the attraction of the screen grid is greater than that of the anode. These secondary electrons subtract from the anode current and augment the screen current. This condition prevails while e_b is within the limits marked X to Y in Fig. 1-45, and between these limits anode current actually decreases as anode voltage is increased. Beyond Y the anode voltage is approaching the screen-grid voltage in value, and the region where the attraction of the screen for secondary electrons is greater than that of the anode is progressively reduced. Consequently more of the secondary electrons return to the anode and screen current begins to fall, as shown in the figure between Y and Z. Within these limits anode current rises rapidly as anode voltage is increased. When the anode voltage is much larger than the screen voltage, virtually all secondary electrons are attracted back to the anode. In addition, many electrons, which formerly were collected by the screen grid, are now travelling with sufficient velocity to pass through the screen grid and reach the anode. Anode current increases and screen current decreases until anode current saturation is reached near W.

Along the portion XY of the static $e_b - i_b$ curve, where an increase in anode voltage causes a decrease in anode current the tetrode is said to have a *negative a-c anode resistance*. This property finds application in certain oscillator circuits, but in most applications is highly undesirable. For this reason the operating point of a tetrode is generally selected to be beyond point Z, where, in addition, the a-c anode resistance is very high. As a result the amplification factor of a tetrode is also very high, and the control grid is a great deal more effective in controlling anode current than it is in the case of a triode.

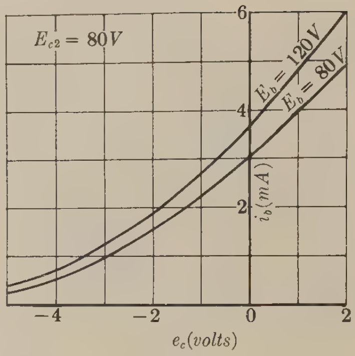


FIG. 1-46 Tetrode static mutual characteristics ($e_c - i_b$ curves).

The amount of secondary emission in a tetrode for given anode and screen potentials depends upon the material of which the anode is made. Carbon and graphite exhibit less secondary emission effect than most materials used in anode construction.

B. The Static Mutual Characteristics or $e_c - i_b$ Curves

To investigate the relationship between the grid voltage and anode current the anode voltage (Fig. 1-43) is first maintained constant at a suitable value and screen voltage E_{c2} at its rated value. The grid voltage is varied in steps and the corresponding values of anode current are recorded. The values of grid voltage are used as abscissae and the values of anode current as ordinates to plot points and so obtain the characteristic curve (Fig. 1-46). The second curve shown is obtained by repeating the procedure with E_b constant at another value and E_{c2} at its rated value. The shape of the static $e_c - i_b$ curve of a tetrode is similar to that for a triode. The screen grid does not prevent the control grid from controlling the anode current. However, because the screen grid is at a high positive potential and is very close to the control grid, the control grid voltage must be very negative before the anode current becomes zero.

1.19 Values of the valve factors for a tetrode. The values of the transconductance (g_m) and anode resistance (r_a) for a tetrode may be obtained from the $e_c - i_b$ curves by the general methods outlined for the triode.

In Fig. 1-47(a), $\triangle KLM$ is drawn along the linear portion of the $e_c - i_b$ curve. When the curve is not entirely linear the tangent is drawn at the required operating point. Thus the hypotenuse of $\triangle KLM$ is the tangent to the curve at the operating point P.

$$\text{Then } g_m = \frac{\Delta i_b}{\Delta e_c} \quad (\text{for } E_b = 90V \text{ and } E_{c2} = 80V)$$
$$= \frac{LK}{LM}$$
$$= \frac{2\times 10^{-3}}{2.8}$$
$$\doteqdot 714\times 10^{-6}$$

Therefore the transconductance is 714 micromhos. Values of g_m for a tetrode range from about 500 to 3500 micromhos.

In Fig. 1-47 (b) the anode resistance for operation at point P may be obtained by using $\triangle KLM$.

$$\text{Then } r_a = \frac{\Delta e_b}{\Delta i_b}$$
$$= \frac{ML}{LK} \quad (\text{for } E_c = -1V \text{ and } E_{c2} = 80V)$$
$$= \frac{112}{2\times 10^{-3}}$$
$$= 56{,}000$$

Therefore the anode resistance is 56,000 ohms. This is considerably larger than that for most triodes. Values of r_a for tetrodes range up to about 1,000,000 ohms if the anode voltage is operated at near saturation voltage.

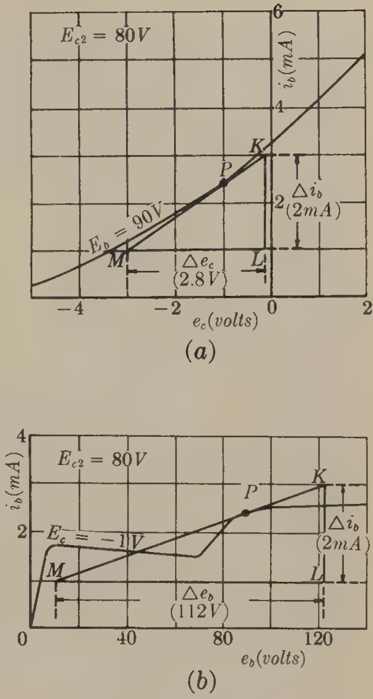


FIG. 1-47 Method of obtaining values of the valve factors for a tetrode.

The amplification factor is obtained from the product of the transconductance and the anode resistance.

$$\text{Then} \quad \mu = g_m \times r_a$$
$$= 714\times 10^{-6}\times 56{,}000 \doteqdot 40$$

Thus the amplification factor is 40. In tetrodes the value of μ ranges from about 40 to 1500.

Because of its low grid-to-anode capacitance and high amplification factor the tetrode is a useful amplifier at the lower radio frequencies.

1.20 The pentode. In the application of tetrodes it is undesirable to operate on the negative resistance portion of the characteristic. Thus it is necessary to operate the anode at a high voltage in order to overcome the effects of secondary emission. In addition, in the amplification process, since a varying voltage is developed at the anode, the anode voltage may become less than the screen voltage when a large varying output voltage is developed across the anode load. When this occurs the undesirable effects of secondary emission are again encountered. As a result the available a-c voltage that is developed at the output of a tetrode is limited. These disadvantages of the tetrode can be eliminated by the insertion of a third grid, the *suppressor grid*, between the anode and the screen grid. With this addition a five-electrode valve, called the *pentode*, is formed. The suppressor grid is constructed of the same material as the other grids but its wires are more widely spaced. Fig. 1-48 shows the construction and circuit symbol of a pentode.

When the suppressor grid is operated at a potential which is negative with respect to the anode, usually by connecting it directly to the cathode, it counteracts but does not prevent secondary emission. Even at low anode voltages, it repels the secondary electrons back to the anode and prevents them from forming a part of the screen current. When operated at cathode potential, the suppressor grid tends to decelerate the electrons from the cathode. However, their velocity is sufficient to carry them into the field of the anode which attracts them through the suppressor.

A transmitter pentode (half actual size) is shown in Fig. 1-49. The parts and assembly steps for a metal-envelope receiver pentode (half actual size) are shown in Fig. 1-50.

1.21 The characteristic curves of a pentode. The static characteristic curves may be obtained using corresponding values of current and voltage provided by the circuit of Fig. 1-51.

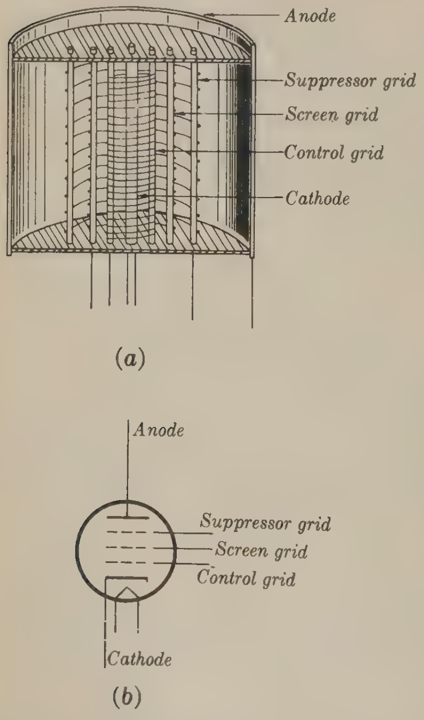


Fig.1-48 (a) Arrangement of electrodes in a pentode (b) Circuit symbol for a pentode with indirectly-heated cathode.

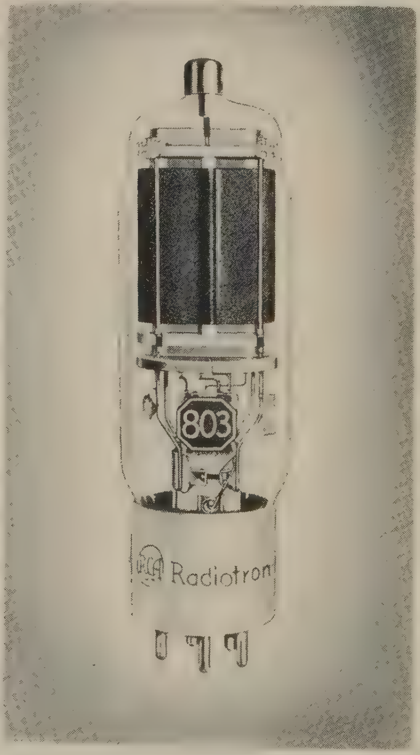
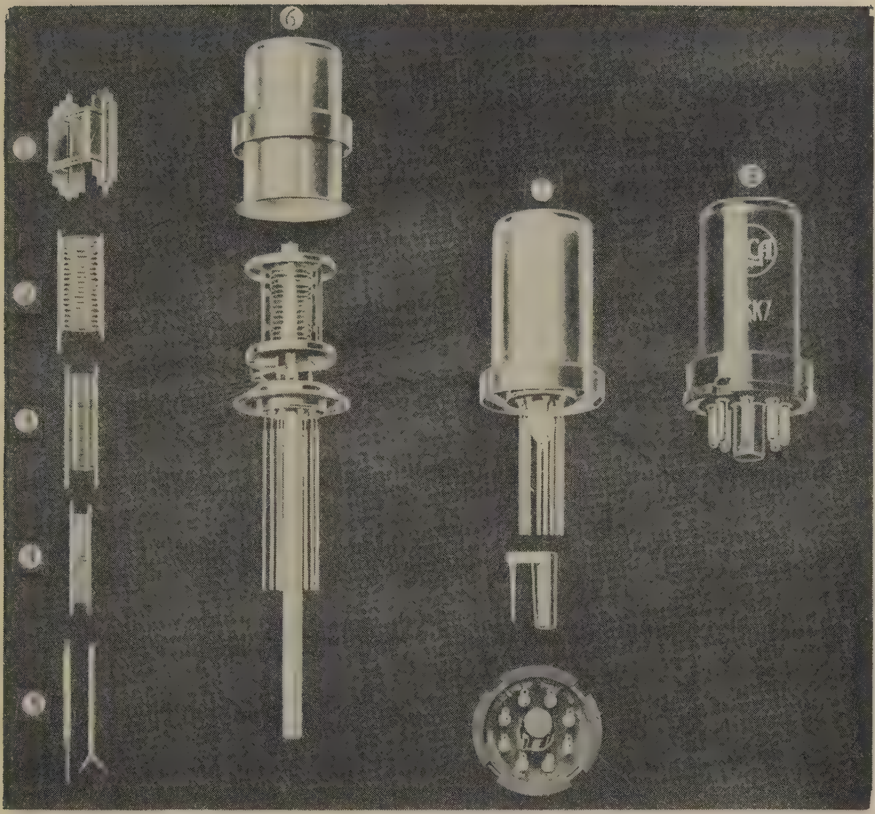


Fig.1-49 A transmitter pentode.



① Anode
② Suppressor grid
③ Screen grid
④ Control grid
⑤ Heater and cathode
⑥ Valve mount, metal shell and collar
⑦ Evacuated envelope, base, shield and base
⑧ Complete valve

FIG. 1-50 Parts of a metal-envelope receiver pentode.

A. The Static Anode Characteristics or $e_b - i_b$ Curves

The static $e_b - i_b$ curve is determined for a constant value of screen-grid voltage. With E_{c2} constant at its rated value and the suppressor grid maintained at zero potential by connecting it to the cathode, the control-grid voltage is first set at 0 volts. Then as the anode voltage is varied in steps from 0 volts up to the anode voltage rating of the valve, the corresponding values of anode current are recorded. The values of anode voltages are used as abscissae and the values of anode current as ordinates to plot

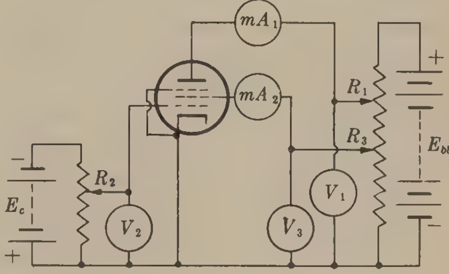


FIG. 1-51 Circuit for obtaining data to plot the characteristic curves of a pentode.

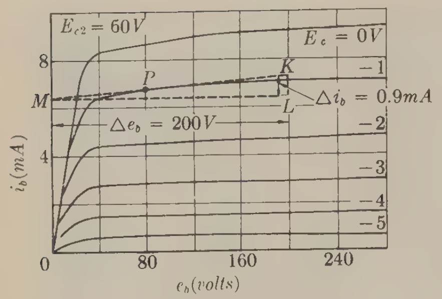


FIG.1-52 Pentode static anode characteristics ($e_b - i_b$ curves).

points and obtain the characteristic curve. By leaving the screen-grid voltage constant, a family of anode characteristic curves (Fig. 1-52) may be obtained by repeating the above procedure for various values of E_c.

From this typical family of curves it is seen that the rise in anode current in a pentode is very rapid. The negative slope portion seen in the tetrode curve has disappeared. The linear portion of the pentode static $e_b - i_b$ curve extends over a much larger range of anode voltage than in the case of the triode or tetrode. Since the linear portion of the pentode static $e_b - i_b$ curve is so nearly horizontal, the anode voltage has practically no control over the anode current and the anode resistance and the amplification factor are very high.

B. The Static Mutual Characteristics or $e_c - i_b$ Curves

The static $e_c - i_b$ curve is determined for constant values of anode and screen-grid voltages. With E_b constant at a suitable value, E_{c2} at its rated value, the control-grid voltage is varied in steps from some negative value beyond cut-off up to zero and the corresponding values of anode current are recorded. The values of grid voltage are used as abscissae and the values of anode current as ordinates to plot the points and so obtain the characteristic curve, as shown in Fig. 1-53. The shape of this curve is similar to the static $e_c - i_b$ curve for a triode or a tetrode.

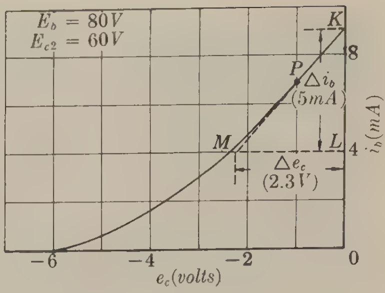


FIG.1-53 Pentode static mutual characteristic ($e_c - i_b$ curve).

1.22 Values of the valve factors for a pentode. The transconductance, g_m, of the pentode for which the $e_c - i_b$ curve is shown in Fig. 1-53 is calculated for the operating point P on the linear portion with the aid of triangle KLM.

$$g_m = \frac{\Delta i_b}{\Delta e_c}$$

$$= \frac{LK}{LM} \quad \text{(for } E_b = 80V \text{ and } E_{c2} = 60V\text{)}$$

$$= \frac{5 \times 10^{-3}}{2.3}$$

$$\doteqdot 2170 \times 10^{-6}$$

Therefore the transconductance is 2170 micromhos. Values of g_m for pentodes are of the same order as those for triodes.

The anode resistance, r_a, of the same pentode is calculated for the operating point P by using triangle KLM along the static $e_b - i_b$ curve of Fig. 1-52.

$$r_a = \frac{\Delta e_b}{\Delta i_b}$$

$$= \frac{ML}{LK} \quad \text{(for } E_c = -1V \text{ and } E_{c2} = 60V\text{)}$$

$$= \frac{200}{0.9 \times 10^{-3}}$$

$$\doteqdot 222{,}000$$

Therefore the anode resistance is 222,000 ohms. For pentodes the values of r_a range from about 100$K\Omega$ to 1$M\Omega$.

The amplification factor is the product

of the transconductance and anode resistance. Since the value of the transconductance is approximately the same as that of a triode or tetrode while the anode resistance is much larger, the amplification factor of a pentode is very large. For the same pentode,

$$\begin{aligned}\mu &= g_m \times r_a\\ &= 2.170 \times 10^{-3} \times 222{,}000\\ &= 480\end{aligned}$$

Therefore the amplification factor is 480. Values of μ for pentodes range from about 100 to 1500.

The grid-to-anode capacitance for a receiver pentode is as low as .003 $\mu\mu F$. For a transmitter pentode, the grid-to-anode capacitance is only 0.15 $\mu\mu F$ as compared with 6 or 7 $\mu\mu F$ for a transmitter triode. The pentode, therefore, is a very useful valve for amplification of voltages at moderate and high frequencies. The table at the bottom of this page gives examples of values of valve factors and interelectrode capacitances for triodes, tetrodes and pentodes of the receiving type.

1.23 The beam power valve. In place of a suppressor grid to control the secondary electrons from the anode of a tetrode, it is possible to obtain the same effect by suitably spacing and shaping the electrodes and inserting two *beam-forming plates*. Fig. 1-54 shows the electrode arrangement and distribution of electrons in such a valve, called a *beam power valve*. This valve is essentially a tetrode with the addition of two beam-forming plates which are connected internally to the cathode so that these plates and the cathode have the same potential. The cathode is flattened as shown so that most of the electrons are emitted in a direction between the beam-forming plates as shown in the figure. Under the attraction of the anode and as a result of the lateral screening of the plates, the electrons emitted from the cathode form two beams between the plates. Each beam covers an angle of

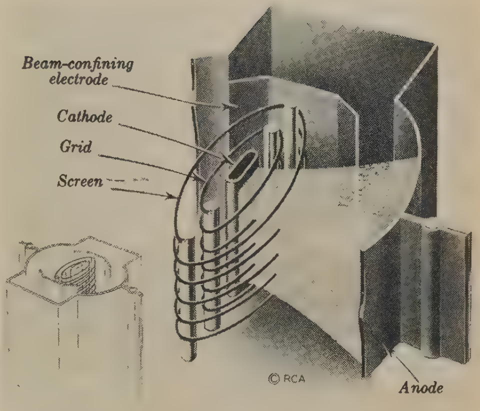


Fig. 1-54 Arrangement of electrodes in a beam power valve.

Valve Factor or Capacitance	Triode	Tetrode	Pentode
g_m in micromhos	1000 to 5000	500 to 3500	1000 to 5000
r_a in Ω	500 to 50,000	50,000 to 200,000	200,000 to 1,000,000
μ	2 to 100	100 to 400	400 to 1500
C_{ga} in $\mu\mu F$	2	.007	.005
C_{gk} in $\mu\mu F$	4	5	5
C_{ak} in $\mu\mu F$	12	10	12

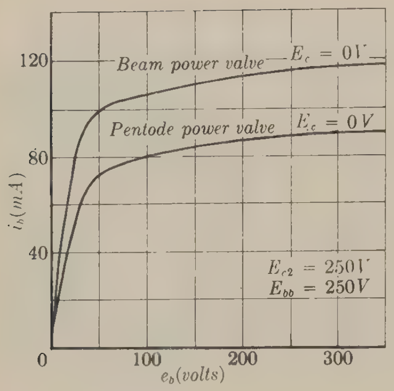


FIG.1-55 Comparative $e_b - i_b$ curves of an a-f power pentode and an a-f beam power valve of the same size.

about 60°. The spacing between the screen grid and anode is greater than in the conventional tetrode.

The screen grid of the beam power valve is maintained at a positive voltage equal to, or greater than, the anode voltage. Therefore, when the electrons pass the screen, they come into a region of relatively low field strength. The electrons slow down and a concentration of electrons is formed in the vicinity of the anode producing a region of minimum potential. As long as the anode potential is greater than this minimum potential any secondary electrons emitted from the anode are repelled back to the anode.

The screen current is prevented from becoming excessive by the position of the screen grid in the valve. Because of its high positive potential the screen grid of a beam power valve would ordinarily draw a larger current than that of a pentode or tetrode. However each turn of wire of the screen grid is shaded from the cathode by a turn of the control grid. In addition, the negative potential of the control grid confines or focuses the electrons into beams between the control-grid wires so that the electrons travel between the screen wires as well. Only a few of the electrons are attracted to the screen. Fig. 1-54 shows this focusing of the electron beam which causes it to pass between the turns of the screen grid. In this way the screen current and associated power loss are reduced to a minimum.

Fig. 1-55 compares the static $e_b - i_b$ curves of an a-f power pentode and an a-f beam power valve of the same size. Because of the effect of the beam focusing which allows few electrons to go astray and because of the low screen current, the anode current in the beam power valve rises to a larger value for a lower anode voltage than in the pentode. Hence the characteristic curve rises more steeply. As a result, the beam power valve has a longer linear range than a pentode. Thus, a beam power valve operated at the same voltages as a conventional pentode or tetrode provides higher anode current and so more power output for a given amount of cathode emission.

In *beam power pentodes* a suppressor grid assists the space charge in preventing secondary electrons from reaching the screen grid.

1.24 The variable-mu pentode. In the valves previously discussed the grid wires are uniformly spaced. However in certain pentodes the grid is constructed with a non-uniform mesh. Fig. 1-56 shows such a grid wound so that the spacing between wires is greater towards the center of the grid than towards the ends. In the end regions the grid exercises good control over the anode current. In the

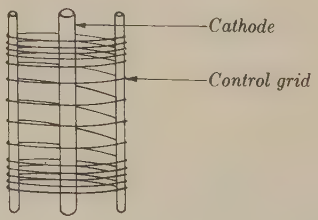


FIG.1-56 Control grid winding in a variable-mu pentode.

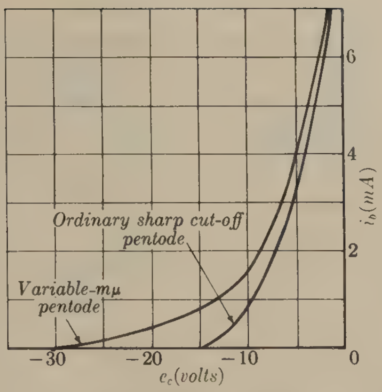


FIG.1-57 Comparative $e_c - i_b$ curves of remote cut-off and sharp cut-off pentodes.

central region almost the same current aflows at all values of grid voltage since the wider spacing of the grid wires reduces the control exercised by this portion of the grid. Cut-off is achieved at low values of negative grid voltage at the end regions where the grid wires are closely spaced. As the grid voltage progressively becomes more negative the current is cut off closer and closer to the centre of the grid. Complete cut-off is finally achieved at a relatively high value of negative grid voltage. The greater the spacing of the grid in the central region, the greater is the cut-off voltage. This type of pentode is called a *remote cut-off pentode*.

The static $e_c - i_b$ curves for remote cut-off and sharp cut-off pentodes are shown in Fig. 1-57. The sharp cut-off pentode characteristic is almost linear and hence the value of transconductance is practically constant for all values of grid voltage. For the remote cut-off pentode, however, the transconductance decreases very gradually from its constant value along the linear portion of the curve and does not finally become zero until the grid bias reaches a large negative value about $-30V$. Thus changing the grid voltage gives a change in the transconductance and hence in the amplification factor of the valve. As the grid bias increases the amplification factor decreases. Because of this property, the valve is also called a *variable-mu pentode* or a *super-control pentode*. It is particularly useful as an r-f amplifier where automatic volume control is desired.

1.25 Multi-purpose valves.

A. Multi-Electrode Valves

Valves containing more than three grids, an anode and a cathode are called *multi-electrode valves*. The anode current is controlled by voltages at various frequencies on several electrodes. Two common types of multi-electrode valves are the *pentagrid converter* and the *pentagrid mixer*, the circuit symbols of which are illustrated in Fig. 1-58 (a) and Fig. 1-58 (b) respectively.

B. Multi-Unit Valves

A valve which contains within a single envelope two or more separate valve units is called a *multi-unit valve*. For example, Fig. 1-59 (a) shows the circuit symbol for a diode-triode. Each section functions in exactly the same way as if each had its own envelope; each section may be connected into two completely different circuits. Such a construction permits a few multi-unit valves to perform the functions of a larger number of single-unit valves. The advantages are a saving in space and costs. Fig. 1-59 (b), (c) and (d) show the circuit symbols for a twin diode, a twin triode and a triode-hexode respectively.

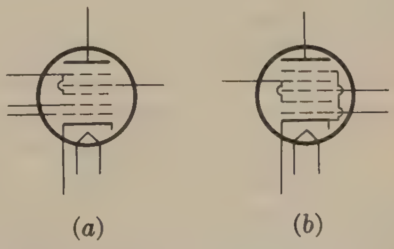

(a) (b)

FIG.1-58 (a) Circuit symbol for a pentagrid converter (b) Circuit symbol for a pentagrid mixer.

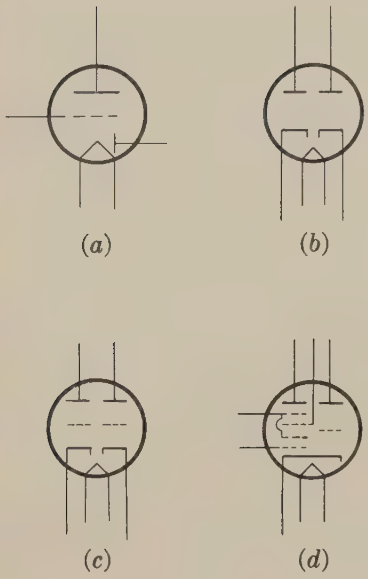


FIG.1-59 Circuit symbols for (a) diode-triode (b) twin diode (c) twin triode (d) triode-hexode.

1.26 Gas-filled valves. A *gas-filled valve*, as its name implies, is one into the envelope of which a small amount of gas such as mercury vapour, neon or argon is introduced after evacuation. The presence of the gas inside the valve produces certain characteristics which make this valve very useful in many circuits. In the valves already discussed the presence of a slight trace of gas such as air has very undesirable effects. In the first place, electrons travelling to the anode strike gas molecules and these molecules break up into positive ions and negative electrons. The positive ions are heavy and move towards the negatively charged cathode. At normal operating voltages the cathode surface is damaged by the bombardment of positive ions. Since most receiver valves contain oxide-coated or thoriated-tungsten cathodes, this positive-ion bombardment destroys the emitting surface. Secondly, in valves containing a control grid another serious effect occurs as a result of the ionization of the gas. Positive ions are attracted to the grid which is normally at a negative voltage. The control grid becomes coated with positive ions. Hence its negative charge is reduced and the anode current increases. This process continues until the grid becomes so positively charged that an excessive anode current flows and the valve is destroyed by the heat developed. Thirdly, the slower-moving positive ions, attracted to the cathode, gather near the cathode at low anode voltages. The effect of the negative electron space charge is nullified and saturation current may be drawn. The valve no longer operates over the linear portion of its characteristic and thus is not fulfilling its function.

However, these effects which are undesirable in high-vacuum valves are put to use in certain gas-filled valves which are usually diodes or triodes. The diodes are operated with either heated or unheated cathodes and are called *hot-cathode gas diodes* or *cold-cathode gas diodes* respectively. The triode gas valve, which is operated with a heated cathode, is called a *thyratron*. The presence of gas in valves is indicated in circuit symbols in this text by diagonal lines as in Fig. 1-60. Another convention is to use a dot in place of the diagonal lines.

Gas valves are also called *soft valves* since the presence of the gas causes the degree of evacuation to be relatively low. When the gas within is ionized these valves give off a luminous glow, the color of which depends upon the kind of gas used. For example, mercury vapour, neon and argon give blue, orange and purple glows respectively.

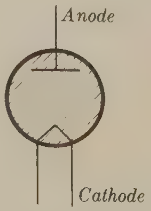


FIG.1-60 Circuit symbol for a hot-cathode gas diode.

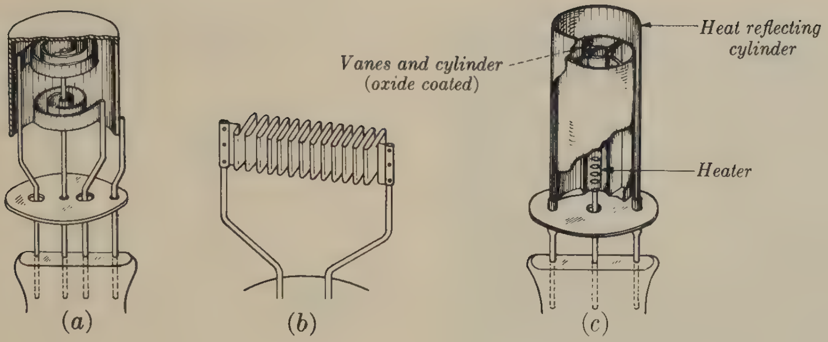


FIG.1-61 Shapes for cathodes of gas valves (a) Spiral (b) Crimped ribbon (c) Concentric hollow cylinders.

A. Hot-Cathode Gas Diode

A hot-cathode gas diode usually contains mercury vapour in equilibrium with a few drops of liquid mercury. When the filament is heated and a positive potential is applied to the anode, an anode current flows. The electrons leaving the cathode encounter and collide with the molecules of the gas. When the anode potential is sufficiently positive the electrons colliding with the gas molecules have sufficient velocity to break up these molecules into positive ions and electrons. The electrons released by the collisions then join the original electron stream and may be capable of liberating other electrons in a similar manner. This process becomes cumulative and is called *ionization* of the gas molecules. The gas diode has molecules, positive ions and free electrons within the envelope. The free electrons, greatly increased in number by the ionization process, travel to the anode. The heavier positive ions drift toward the cathode.

When the positive ions reach the space charge surrounding the cathode, the free electrons in the space charge combine with them to re-form gas molecules. The space charge is almost completely neutralized and the full cathode emission can flow to the anode with a much lower anode voltage than in a high-vacuum diode. This lower voltage is usually of the order of $10-15V$. There are several advantages resulting from the neutralization of the space charge.

(i) Indirectly-heated oxide-coated cathodes which are more efficient than directly-heated types may be used. This is permissible because, with the low anode voltage, the positive ions are moving too slowly to cause disintegration of the cathode.

(ii) It is possible to use cathodes which are constructed with various shapes such as spirals, Fig.1-61(a), crimped ribbons, Fig.1-61(b), and concentric hollow cylinders, Fig.1-61 (c), coated on the inside and outside to give a large emitting surface. The positive ions can penetrate into the hollows and enclosures to neutralize the space charge.

(iii) The spacing between anode and cathode may be increased, thus allowing for larger cathode structures and at the same time permitting a reasonably large negative anode voltage.

The energy required to dislodge electrons from their orbits to produce ionization is supplied by means of the voltage between the anode and cathode. There is a certain anode voltage at which ionization occurs for each gas-filled valve.

The anode potential at which ionization commences is known as the *ionization potential, striking potential* or *firing point.* When the striking potential is reached the entire cumulative process of ionization is virtually instantaneous and a large current flows through the valve. After ionization has started, the ionization can be maintained by an anode potential which is considerably lower than the striking potential. However, a minimum voltage exists which is needed to maintain the ionization. If the anode voltage falls below this value, the gas de-ionizes and the anode current decreases suddenly to a very small value. This critical voltage is known as the *de-ionizing potential* or *extinction potential.*

A gas diode can be used as an electronic switch. At the striking potential the valve ionizes and permits a large current to flow, then at the extinction potential the valve de-ionizes blocking current flow. This valve in a circuit presents a very low resistance when ionized but an almost infinite resistance when de-ionized.

An anode characteristic for one hot-cathode gas diode is shown in Fig. 1-62. The change in anode current for a change in anode voltage is very small for anode voltages below the ionization potential. At the ionizing potential the anode current increases very suddenly to its high saturation value. If the anode voltage is increased beyond striking voltage there is little further change in current. However, the velocities of the positive ions are increased and their bombardment may destroy the cathode. At anode voltages below $22V$ for this particular valve the velocities of the positive ions are not sufficient to cause damage. Therefore when a high voltage is used a resistor must be connected in series with the gas diode (Fig. 1-63) in order to limit the current to some value less than saturation. Most of the applied high voltage appears across the resistor and a small voltage about equal to the ionizing potential appears across the valve.

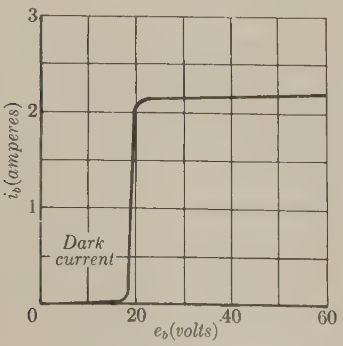


FIG.1-62 An anode characteristic for a hot-cathode gas diode.

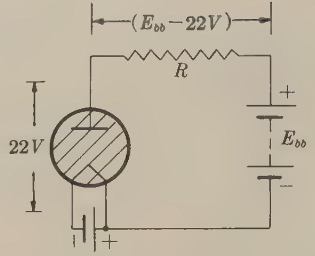


FIG.1-63 Circuit arrangement of a hot-cathode gas diode.

Example. If the saturation current of a mercury-vapour diode is $2A$ and the ionization potential is $20V$, calculate the smallest possible value of a load resistor that could be used in series with the diode, when $E_{bb} = 200V$.

Voltage across the valve $= 20V$

Voltage across the resistor $= 180V$

Maximum current in the resistor $= 2A$

$$\text{Minimum value of } R = \frac{180}{2} = 90$$

Therefore the minimum value of the resistance is 90 ohms.

The ionization potential depends on the pressure of the gas. If the pressure were lowered in a particular valve a higher anode voltage would be required to cause ionization. At this high voltage, positive-ion bombardment might damage the cathode. If the pressure of the gas were raised, electron emission would be reduced and again a higher anode potential would be required for ionization, with consequent damage to the cathode. In mercury-vapour diodes the pressure is usually within the range 1×10^{-6} to 30×10^{-6} mm. of mercury. Since the pressure of the gas depends on the temperature, the temperature must remain within certain limits in order that the pressure may be correct. Since the required temperature for a high-power valve is considerably above room temperature, the valve should be heated before the anode voltage is applied. This is usually accomplished by applying the filament voltage first and allowing the heated filament to warm up the valve. By the use of a time-delay relay the anode voltage can be automatically applied after a suitable warming-up period.

The chief advantage of a gas diode is that current as high as saturation current may be obtained for anode voltages as low as 15 to 20V. The valve's internal resistance and the valve's power loss are low. Gas diodes are used primarily as rectifier valves. Fig. 1-64 shows a mercury-vapour diode rectifier. This valve operates with a filament voltage of 5V and a filament current of 7.5A. The anode current may be as high as 5A. The ionization potential is about 10 volts. The filament must be heated for 30 seconds to heat the mercury vapour to its operating temperature before the anode voltage is applied.

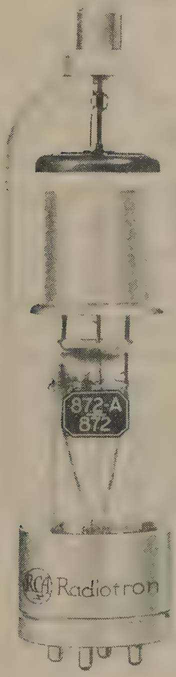


FIG. 1-64 A mercury-vapour diode rectifier.

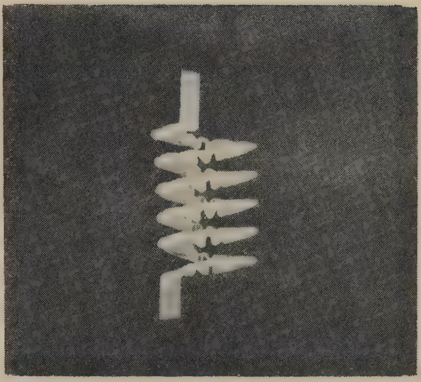


FIG. 1-65 Cathode of mercury-vapour diode consisting of zig-zag ribbon.

For low-power applications the construction of the anode and cathode is similar to that used in high-vacuum diodes. If the gas valve is required to conduct large currents such as the one described above, the cathode consists of a heavy coiled or zig-zag ribbon to produce the required heavy emission. Such a cathode is shown in Fig. 1-65. To increase the emission efficiency of the cathode, the heat it radiates is retained by a cylinder surrounding the cathode as in Fig. 1-64. The inside of the cylinder which is oxide-coated is heated by this radiation to a temperature sufficient to cause emission. Thus the heat radiated by the cathode is not wasted. With this construction, the emission efficiency may be as high as 1.25A per sq. cm. per watt compared to the maximum of 125mA per sq. cm. per watt for the construction of oxide-coated cathodes described in section 1.7. Such a cathode is said to have a *high thermal efficiency*. The anode is a small, flat, inverted cup of metal at the top of the valve.

A disadvantage of a gas diode is the possibility of conduction in the reverse direction through the valve when the anode is at a negative potential with respect to the cathode. In some circuits this voltage in the reverse direction reaches a high value but for proper operation the diode must not conduct. If this inverse voltage is too large, the free electrons which are always present to some extent in an un-ionized gas are accelerated toward the cathode with sufficient velocity to cause a sudden ionization of the gas. A large destructive current then flows through the valve from anode to cathode. This flow of current in the reverse direction through a valve is called *arcing* or *arc-back*. Hence a valve with a given spacing between anode and cathode can be used in circuits with a-c voltages provided the peak negative voltage or *peak inverse voltage* is less than the critical arcing voltage. For the valve shown in Fig. 1-64, the peak inverse voltage allowable is 10,000 volts provided the temperature outside the valve, or *ambient* temperature, remains within the range 20 to 60°C. Should the ambient temperature reach 70°C the gas pressure inside the valve increases and the maximum peak inverse voltage is only 5,000V.

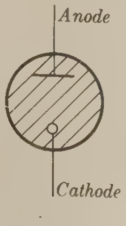


FIG.1-66 Circuit symbol for a cold-cathode gas diode.

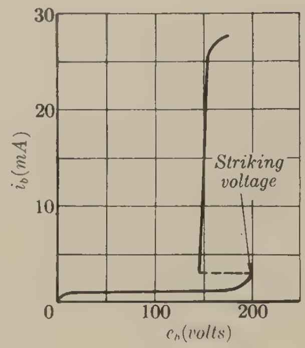


FIG.1-67 An e_b-i_b curve of a cold-cathode gas diode.

B. Cold-Cathode Gas Diodes

A cold-cathode gas valve contains an unheated cathode and a small amount of inert gas such as neon or argon. Valves employing neon are referred to as *neon valves*. Due to the action of light or other stray radiation some free electrons exist in the gas. When a positive voltage is applied to the anode these electrons are attracted to the anode and a small current flows. When the anode voltage increases to a certain positive value the velocity of these electrons is sufficient to cause ionization of the gas. Positive ions drawn to the cathode form a large positive space charge surrounding the cathode. Electrons are pulled from the cathode by the large electrostatic field created between the cathode and the positive space charge. It is possible that some secondary emission due to positive-ion bombardment may take place from the cathode. As a result when ionization occurs a large anode current suddenly flows. The gas is said to *break down* at this value of anode voltage which is, as before, called the *striking voltage*.

The circuit symbol of a cold-cathode gas diode is shown in Fig. 1-66. The e_b-i_b curve of such a valve is shown in Fig. 1-67 and may be obtained using the circuit shown in Fig. 1-68. The resistor R is included in the circuit to limit the current through the valve to a

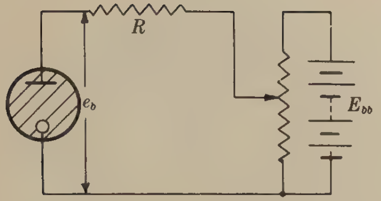


FIG.1-68 Circuit for obtaining data to plot the $e_b - i_b$ curve of a cold-cathode gas diode.

safe value. The value of the anode current at anode voltages lower than the breakdown voltage is of the order of microamperes. At the striking voltage the current suddenly becomes larger and is limited only by the resistance R. At the same time the voltage across the valve drops to its operating value. This is a steady voltage which is maintained as long as the gas is ionized. The striking potential is about one-third higher than the operating voltage. After striking, the current through R is just sufficient to drop the voltage across the valve to the operating value. For a neon valve this is about 150 volts. Cold-cathode gas diodes are also constructed with operating voltages of 75 and 105 volts. If the applied voltage is increased the valve current increases. However, the p.d. across the resistance also increases, leaving the voltage across the valve the same as before. Hence the valve current may vary from 5 to $25mA$ but the anode voltage remains constant at the operating voltage. Because of constant anode voltage during operation, cold-cathode gas valves are used as *voltage regulators.* Fig. 1-69 shows a cold-cathode gas valve. The valve shown has a breakdown voltage of about $180V$ and its operating voltage is $150V$. The permissible variation in current is from 5 to $30mA$.

The breakdown voltage depends on the kind of gas used, the pressure of the gas and the spacing between the anode and cathode. After breakdown the valve gives off a luminous glow.

1.27 The thyratron. The *thyratron* is a hot-cathode mercury-vapour triode.

With control grid operated at zero potential, the mercury vapour becomes ionized and saturation current flows at an anode voltage of about $22V$. If the grid voltage is made negative, a higher anode voltage is required to produce ionization. This effect occurs because the negative grid voltage decreases the velocities of the electrons. Ionization occurs only when the anode voltage is increased to the value at which the electron velocities are sufficiently great to produce ionization. The more negative the grid voltage, the higher is the ionization potential. Thus, the control-grid voltage determines the potential of the anode at which ionization occurs.

Once saturation current flows, the control grid has no further control over the anode current. This situation results from the formation of a sheath of positive ions around the control grid, which nullifies the negative grid voltage. When

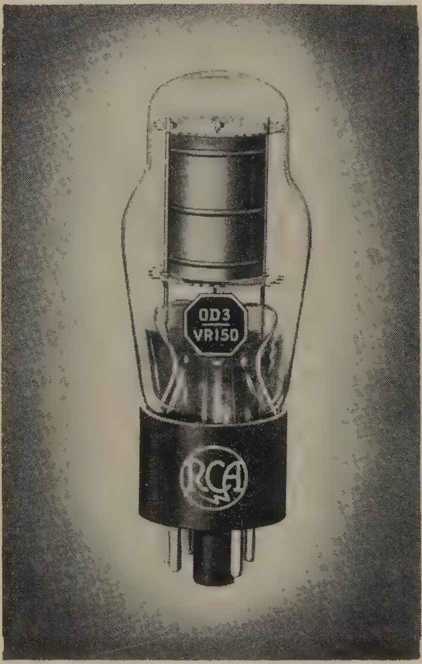


FIG.1-69 A cold-cathode gas valve.

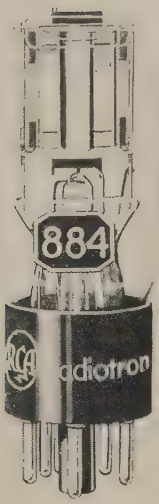


FIG.1-70 A thyratron.

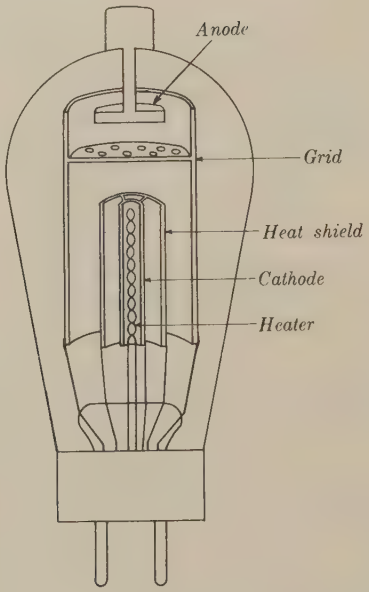


FIG.1-72 Arrangement of electrodes in one type of thyratron.

this happens, a small grid current may flow. Its value is reduced by a series resistor in the grid circuit.

Increasing the negative voltage on the control grid does not de-ionize the valve. In order to cut off the thyratron, the anode voltage must be reduced to a value lower than that necessary to maintain ionization.

One type of thyratron construction is shown in Fig. 1-70. The construction of this thyratron is similar to that of a high-vacuum triode. An oxide-coated cathode is surrounded by a control grid in the form of a spiral which in turn is enclosed by a cylindrical anode. The circuit symbol of a thyratron is shown in Fig. 1-71.

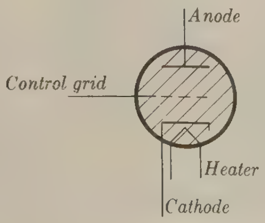


FIG.1-71 Circuit symbol for a thyratron.

Another type of construction is shown in Fig. 1-72. The anode of this thyratron is a small circular plate made of carbonized nickel to increase its ability to radiate heat. Since the voltage across the valve is always less than about 22 V due to the action of a current-limiting resistor in series with the valve, the cathode is of the oxide-coated type. At such a low voltage, positive-ion bombardment is not injurious to the oxide-coating. The cathode is the same as the type described in the section on hot-cathode gas-filled diodes. The controgrid consists of a metal cylinder surl rounding both the anode and cathode. A small horizontal baffle with one or more apertures fits closely into the cylinder between the anode and cathode and is also part of the grid. This cylinder shields the action of the valve from the effect of charges which may collect on the glass envelope. These charges are then prevented from interfering with the control of the grid over the striking potential of the anode. The baffle also repels stray electrons away from the

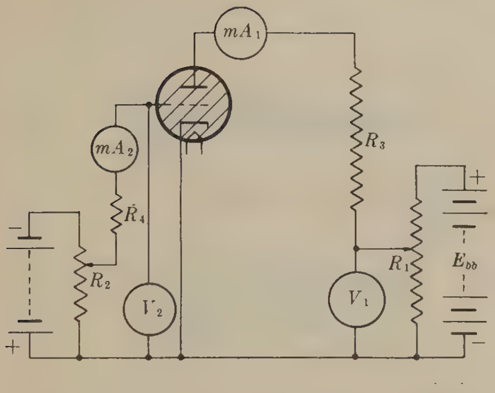


FIG. 1-73 Circuit for obtaining data from which to plot the $e_b - i_b$ curve of a thyratron.

anode, keeping the anode current negligible for anode voltages less than the striking voltage.

The $e_b - i_b$ curves of a thyratron may be obtained using the circuit shown in Fig. 1-73. The grid potential is set at some negative value by means of the potentiometer R_2. The resistance R_4 limits the grid current after ionization. The grid voltage E_c is measured by the voltmeter V_2 and the grid current by the milliammeter mA_2. The anode voltage e_b measured by the voltmeter V_1 is gradually increased from zero by the potentiometer R_1. The discharge current i_b is observed by means of the milliammeter mA_1. R_3 is a series resistor which limits the voltage across the thyratron during ionization to a safe value of about $22V$. Then, by means of R_1, the anode voltage is reduced to zero again, and the process repeated for a different value of control-grid voltage. The resulting $e_b - i_b$ characteristic curve is shown in Fig. 1-74.

When $E_c = 0V$ the thyratron acts as a gas diode. The striking potential is about 22 volts and when this voltage is reached the current tends to rise rapidly to saturation. However the limiting resistor R_3 sets the value to which the anode current rises. As the voltage is further increased the anode current rises linearly in a manner determined by the value of R_3. This happens because the resistance of the valve when it is ionized is very small and hence R_3 alone controls the current. With $E_c = 0V$ a characteristic is drawn for the range of anode voltages from 0-250V and load resistance R_3. When E_c has some other negative value the anode voltage necessary for ionization is increased, but when the valve ionizes the current rises immediately up to the same $E_c = 0V$ characteristic. When $E_c = -3V$ the striking voltage is $42V$ and when $E_c = -15V$ the striking voltage is $220V$. After striking the $e_b - i_b$ characteristic is a straight line.

There is no useful $e_c - i_b$ curve since the grid has no control over the anode current during ionization. However, another useful curve, in which the striking anode voltage is plotted against the control grid voltage, is shown in Fig. 1-75. This is

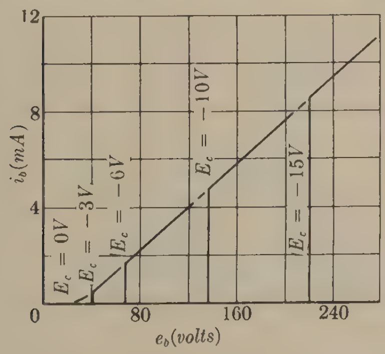


FIG. 1-74 Thyratron $e_b - i_b$ curve.

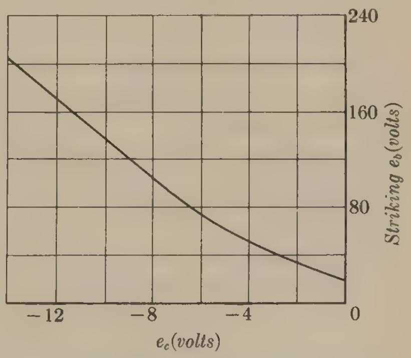


FIG. 1-75 Graph of striking voltage vs control-grid voltage.

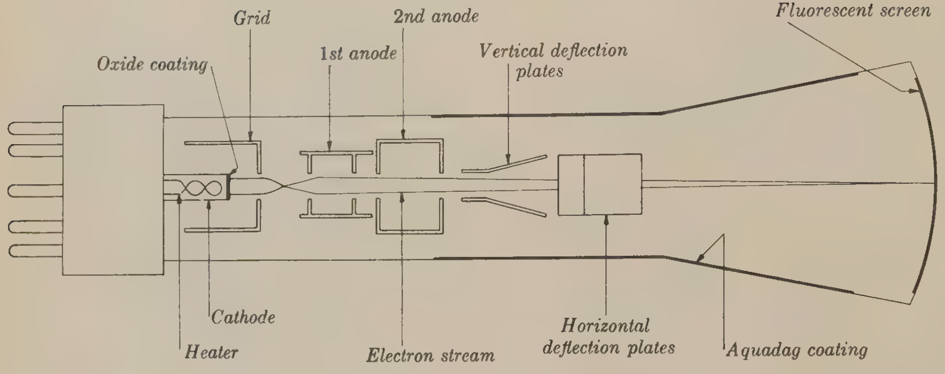


FIG. 1-76 Arrangement of electrodes in an electrostatic cathode-ray tube.

useful in determining the grid striking voltage for any value of E_{bb}.

The thyratron acts as an open circuit before discharge and as a very low resistance during discharge. The voltage of the control grid changes the thyratron from the first condition to the second but it is the anode voltage which must be reduced in order to return the thyratron to its non-conducting state. A thyratron is used as an electronic switch which acts almost instantaneously. Only about 10 microseconds are required for the discharge to reach full value after striking. About 100 to 1000 microseconds are required for the gas to de-ionize depending on the kind of gas used. The de-ionization times are less for valves containing inert gases such as neon or argon. These gases permit normal operation over a wider range of temperature than mercury vapour. However, a mercury-vapour valve has a lower resistance when conducting and can also withstand a higher peak inverse voltage.

The thyratron shown in Fig. 1-70 has the following ratings:

Heater voltage 6.3 V
Heater current 0.6 A
Voltage drop across the valve 10 to 24 V
Max. peak inverse voltage 350 V
Max. anode current 300 mA
Temperature range 35 to 75°C

1.28 Cathode-ray tube. The *cathode-ray tube* is a type of valve in which electrons emitted from a cathode are moved at a high velocity, formed into a narrow beam and then allowed to strike a chemically prepared screen which fluoresces when bombarded by electrons travelling at high speed. The cathode-ray tube is the fundamental device used for visual indication in radar, television and general radio repair and research work. Because the electrons in the beam are very light, the beam may be deflected very rapidly. This property enables the cathode-ray tube to be used at frequencies much higher than those possible with any other type of indicating or measuring device.

Cathode-ray tubes can be divided into two general types, *electrostatic* and *electromagnetic*, depending upon the method of focusing and deflecting the electron beam.

A. *The Electrostatic Cathode-Ray Tube*

In this type of cathode-ray tube the focusing and deflection of the electron beam is done by means of electrostatic fields. The components of one electrostatic cathode-ray tube are illustrated in Fig. 1-76. Fig. 1-77 shows a cut-away photograph of an electrostatic cathode-ray tube which has a screen 5 inches in diameter.

1. CATHODE.

The cathode produces the stream of

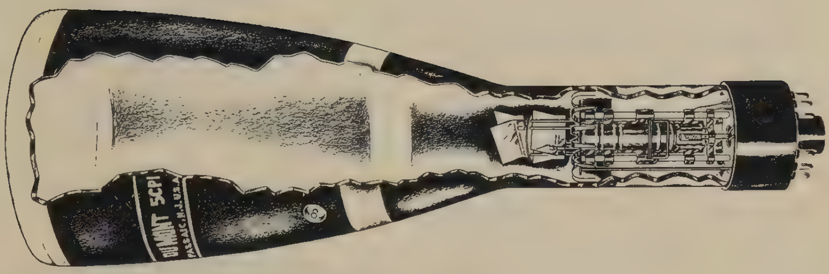


Fig.1-77 A 5-inch electrostatic cathode-ray tube.

electrons. It is usually indirectly heated and consists of a nickel cylinder or sleeve, capped with oxide coating only at the end. This ensures emission in the desired direction only. The heater is a tungsten-wire filament which is wound in the form of a non-inductive double spiral. This construction tends to cause any magnetic field which might affect the electron beam to be cancelled. The magnetic field of half the winding is cancelled by the equal but opposite magnetic field of the other half. The heater coil is insulated and inserted into the cathode sleeve. To provide good heat conduction the insulating material of the heater coil makes contact with the nickel sleeve.

2. Grid.

The grid is used to control the *intensity* of the beam. It is a cylindrical sleeve completely enclosing the cathode except for a small opening opposite the oxide coating. This opening permits the passage of a narrow beam of electrons which are travelling toward the screen. The potential of the grid is usually maintained negative with respect to the cathode. The attraction of the electrons from cathode to anode may be lessened by an increased negative voltage on the grid. The control of the grid potential gives control of the number of electrons in the beam and hence permits control of the brightness of the spot on the screen.

In addition a focusing action takes place as the electrons pass through the electrostatic field between the grid and cathode. This field is represented in Fig. 1-78. An electron emitted in the direction KA is repelled by the electrostatic field which tends to move the electron in a downwards direction. As a result, the electron is deflected and fol-

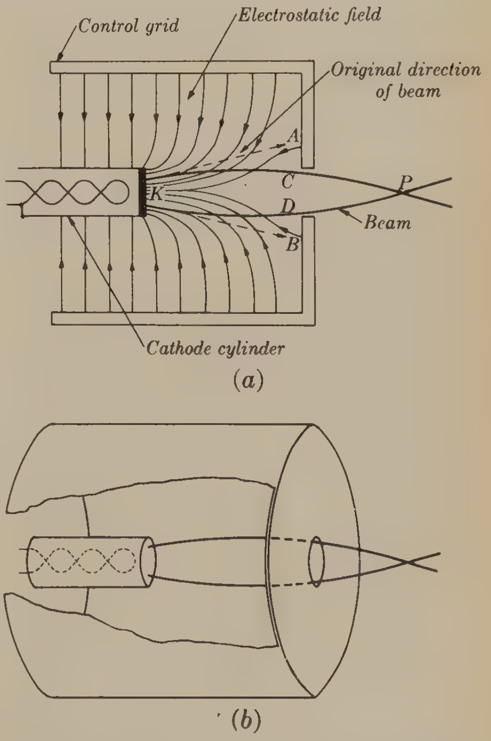


Fig.1-78 Focusing action as a result of electrostatic field between control grid and cathode.

lows the path KCP through the aperture. Similarly, an electron emitted in the direction KB is deflected upwards to travel along the line KDP through P The electrons emitted in the directions between KA and KB are deflected to a lesser extent and also tend to go through the point P. Thus the grid concentrates the electrons into a narrow beam and causes the beam to have a cross-over point in front of the aperture in the control grid.

3. Focusing Anode and Accelerating Anode.

Although the grid helps to narrow the electron beam, it is not capable of sufficient focusing action. The beam is focused at the cross-over point a short distance from the grid aperture but beyond that point it tends to diverge again. The function of the first and second anodes is to focus the electron beam again into a well defined point at the screen and to provide acceleration of the electrons in the beam so that they have sufficient velocity to cause fluorescence when they strike the screen (Fig. 1-76).

The *first anode* is operated at a positive potential of about 300-500 volts with respect to the cathode. The first anode attracts the electrons from the cathode and these form a beam which enters the central regions of the two anodes. Both the first and second anodes are cylindrical and hollow and contain diaphragms with very small apertures at the center. Stray electrons are collected by these diaphragms and only those electrons which are moving straight down the tube near the axis are allowed to pass.

The *second anode*, placed next to the first anode, is slightly larger in diameter. This second anode is maintained at a very high positive potential with respect to the cathode, that is, about four or five times that of the first anode. This potential difference results in a large field between the first and second anodes which causes a large acceleration of the electrons in the beam. For this reason the second anode is sometimes called the *accelerating anode.*

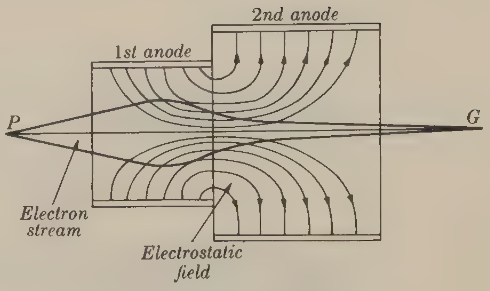
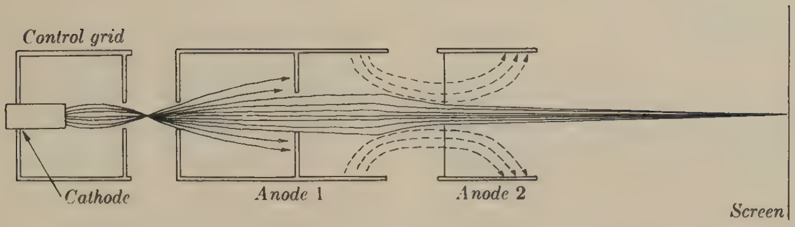


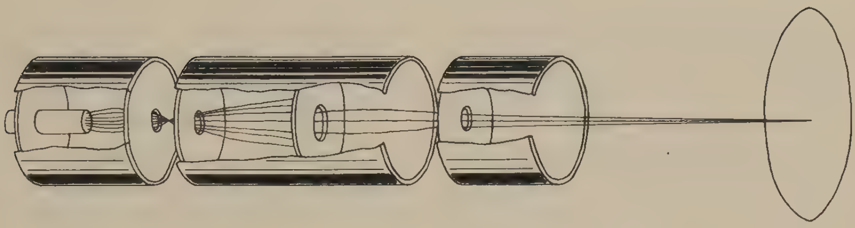
Fig. 1-79 Electrostatic field between first and second anode producing focusing action.

The electrostatic field between the two anodes produces the focusing action on the electron stream. The shape of the field is shown in Fig. 1-79. An electron, if placed at a point in this field, would move in the direction of the lines of force shown in the diagram. Hence, the electrons begin to converge rapidly in the region of the first anode. As they enter the second anode, they are greatly accelerated by the high potential. The lines of force also change direction so that as the electrons pass quickly through the second anode they tend to converge less rapidly. Thus after the beam has passed through anode 2, a second cross-over point occurs. This point of convergence is at G. The potential difference between the two anodes determines the strength of the field and hence the position of G. In practice the second anode is held at a constant potential while the potential of the first anode is varied so that the focal point is at the fluorescent screen. Since the potential of the first anode is varied to focus the beam it is sometimes called the *focusing anode.*

The electrodes of the cathode-ray tube which produce, accelerate and focus the electron beam make up what is known as the *electron gun.* Thus the electron gun



(a)



(b)

FIG.1-80 The electron gun of an electrostatic cathode-ray tube.

consists of the cathode, control grid and two anodes (Fig. 1-80).

In the construction of the electron gun of some cathode-ray tubes more than two anodes or physically split anodes are used. Fig. 1-81 shows the cathode arrangement in a 7-inch electrostatic c.r.t. The cylindrical anode (1) adjacent to the control grid is an accelerating anode. The circular disc (2) which follows is the focusing anode. The short cylinder and attached circular disc (3) is electrically connected to the first cylindrical anode as shown. Together these two anodes become the accelerating anode structure. In this case the accelerating anode may be considered as a split-anode structure. Besides providing additional electron acceleration this arrangement isolates the focusing anode from the control grid and so effectively reduces any interaction between intensity and focus control adjustments. This placement of the focusing anode permits use of a larger aperture reducing current drawn from the beam by this anode and so reducing its effect on the intensity of the beam.

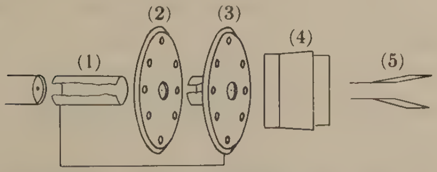


FIG.1-81 Anode arrangement in a 7-inch electrostatic c.r.t.

4. DEFLECTION PLATES.

The *deflection plates* provide a means of deflecting the electron beam as a whole either vertically or horizontally. Two pairs of plates (Fig. 1-76) are mounted in the neck of the tube and situated so that the electron beam passes between each pair. The second pair of plates is arranged at right angles to the first pair.

Since the electrons which make up the electron beam are negative particles of electricity, their movement and hence the deflection of the beam as a whole is governed by the law of attraction and repulsion of charges, that is, like charges repel each other while unlike charges attract. Thus a positively charged deflecting plate attracts the beam while a negatively charged deflecting plate repels it.

To understand fully the action of the

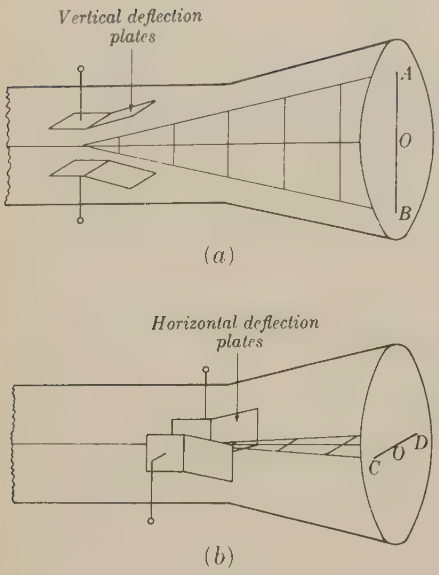


Fig. 1-82 Deflection plates of an electrostatic c.r.t.

deflection plates consider first Fig. 1-82 (a). After leaving the second anode, the beam passes between the *vertical-deflection plates.* With no voltage applied to these plates and with the intensity and focus controls adjusted for optimum performance the beam causes a brilliant, well defined spot of light to appear on the screen at O. When a positive potential is applied to the upper plate and a negative potential to the lower plate an electrostatic field exists between these plates. When electrons of the beam pass through the electrostatic field between the plates they tend to drift towards the positive plate. However, the speed of the electrons is sufficient to allow them to pass through between the plates and the spot on the screen appears at some point A above O. When the potentials applied to these plates are reversed, the beam is deflected vertically downward and the spot appears at B below O.

After leaving the vertical-deflection plates, the beam passes between the *horizontal-deflection plates*, Fig. 1-82 (b). Horizontal deflection is effected in the same manner as is vertical deflection. By changing the horizontal deflection voltage, the spot on the screen can be made to move from O to any point on the line CD. The vertical-deflection plates are sometimes called the *Y-plates* and the horizontal-deflection plates the *X-plates*; these alternative names are given because the direction of the motion of the beam across the screen caused by each set of plates in turn corresponds to the co-ordinate axes of a graph placed on the screen.

The amount the electron beam is deflected from its normal position depends on the velocity of the beam and the strength of the deflecting field. The distance the spot is moved, in millimeters, across the screen by a difference in potential of one volt d-c between the deflecting plates is called the *deflection sensitivity* of the tube. The deflection sensitivity may be increased by increasing the distance from the deflection plates to the screen but this would increase the physical length of the tube. In addition the deflection sensitivity may be increased by increasing the length of the deflection plates. The field is effective on the beam for a longer period of time so that greater deflection of the beam can be caused. In this case the ends of the plates nearest the screen must be bent away from the tube axis so that the electron beam will not strike the plates when it is deflected. Typical deflection sensitivities range from .034 to .34mm. per volt d-c. Amplifier circuits can be used to increase the amplitude of the deflection voltages in order to increase the deflection for a given applied voltage.

Most of the important applications of the electrostatic cathode-ray tube require that voltages be applied to each pair of deflection plates at the same time. The electron beam is then continually acted upon by two forces at right angles

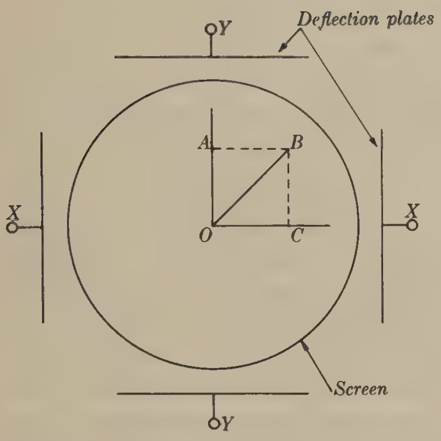


FIG.1-83 Position of spot on screen under action of deflecting voltages applied across deflecting plates.

to each other. Consider the effect of a d-c positive voltage applied to the upper *Y*-plate. The spot moves from its normal position *O* to the point *A* as shown in Fig. 1-83. If this voltage remains constant, and a positive voltage is applied to the right *X*-plate, the spot moves from *A* to *B*. If the order in which these voltages are applied is reversed, the spot moves from *O* to *C* and then from *C* to *B*. If each of the voltages is continuously increased from zero to its final value at the same time, the spot moves along the line *OB*. Thus the resultant of the two forces tending to move the spot to *A* and to *C* is the force which moves the spot to *B*.

When an alternating voltage of low frequency is applied to a pair of deflection plates, the spot moves back and forth across the screen. If the frequency of this applied voltage is increased beyond a certain limit it becomes impossible for the eye to follow the motion of the spot. Due to persistence of vision the path of the spot appears as an illuminated line. In most cases this frequency at which the moving spot produces a continuous line or trace on the screen is about 20 c/s.

Since the electrons in the beam have negligible mass and hence negligible inertia, the beam can be moved very rapidly. The beam responds to deflecting voltages at frequencies ranging up to several hundred megacycles per second.

The deflection plates are usually connected to, and hence have the same potential as, the second anode which is operated at a potential of several thousand volts positive with respect to the cathode. If the cathode is connected to ground this arrangement is dangerous to personnel since the connections to the deflection plates are handled in experimental work. Therefore it is more usual to operate the second anode and the deflection plates at ground potential. In order that the second anode be positive to the cathode, the cathode must be operated at a high negative potential.

5. SCREEN.

The screen at the viewing end of the tube consists of a thin interior coating of fluorescent material Such substances as willemite (zinc orthoscilicate) emit light when struck by fast moving electrons. Different substances give different colours; willemite gives predominantly green light; zinc beryllium silicate, yellow light; zinc oxide, blue light; a mixture of zinc sulphide and cadmium sulphide, white light. The duration of afterglow or phosphorescence called the *persistence* varies with the material. Tubes used for test purposes generally use a substance having a short persistence while tubes used to observe periodic phenomena which occur at a low repetition rate require a fluorescent material having a long persistence.

6. ENVELOPE.

The glass envelope enlarges toward the screen end which may have a diameter up to the order of 24 inches.

Nearly all cathode-ray tubes, whether electrostatic or electromagnetic, are coated near the screen on the inside of the glass envelope with *aquadag* which is a

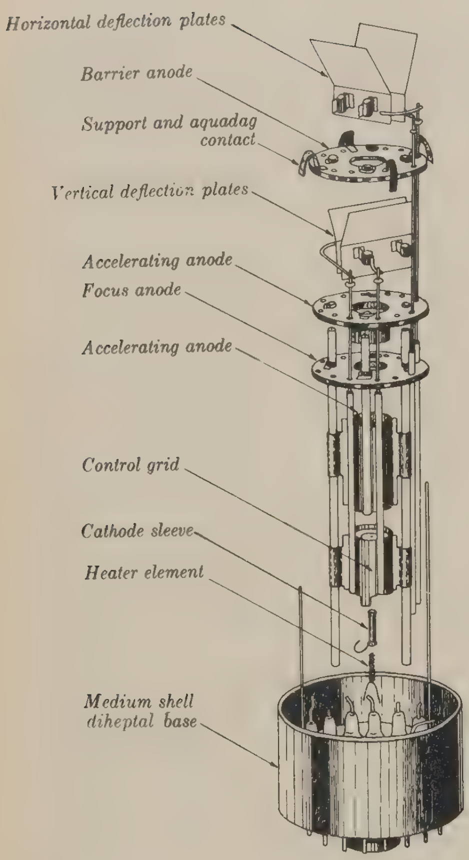

Horizontal deflection plates
Barrier anode
Support and aquadag contact
Vertical deflection plates
Accelerating anode
Focus anode
Accelerating anode
Control grid
Cathode sleeve
Heater element
Medium shell diheptal base

Fig. 1-84 Arrangement of electrodes in a 7-inch electrostatic cathode-ray tube.

form of graphite. When the high velocity electrons in the beam strike the fluorescent screen, secondary electrons are emitted. The main function of the aquadag is to collect the secondary electrons and conduct them back to the anode circuit. This prevents any accumulation of static charge on the viewing end of the tube, thereby eliminating the danger of shock. Stray charges which may collect on the walls of the envelope and affect the direction of the electron beam are also conducted back to the anode circuit. Another function of this graphite coating is to provide shielding for the electron beam. In some cathode-ray tubes, the aquadag is connected to a high positive potential to act as an accelerating electrode.

Fig. 1-84 shows a 7-inch cathode-ray tube structure using electrostatic deflection.

B. The Electromagnetic Cathode-Ray Tube

The electromagnetic cathode-ray tube uses an electromagnetic method of focusing and deflecting the electron beam. In all other respects it differs little from the electrostatic type. Electromagnetic focusing has the advantage of giving a sharper picture on the screen. Electromagnetic deflection also has advantages where a rotating sweep is used.

The elements of an electromagnetic cathode-ray tube are shown in Fig. 1-85. The cathode and grid structures are similar to those used in the electrostatic type. The first anode is cylindrical, with a small aperture in one end to define the beam. Its potential, which is positive with respect to the cathode, gives the electrons their initial acceleration. Further acceleration is obtained by the potential on the aquadag coating, which is still more positive with respect to the cathode. The screen is a fluorescent coating at the end of the tube on the inside of the envelope. The tube shape is different from that of the electrostatic type and the tube is usually shorter. This shorter length of the tube requires the anode voltages to be higher in order to produce the acceleration of the electrons necessary to obtain suitable fluorescence.

1. Focusing.

To understand the action of magnetic focusing it is well to review the effect of a magnetic field upon a current-carrying conductor. An electric current in a conductor consists of a stream of electrons and a magnetic field exists about the conductor. If such a current-carrying conductor is placed at right angles to a uniform magnetic field, interaction of the

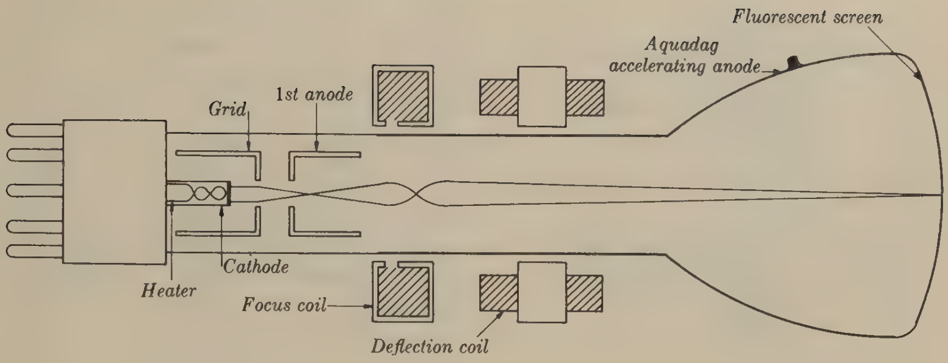
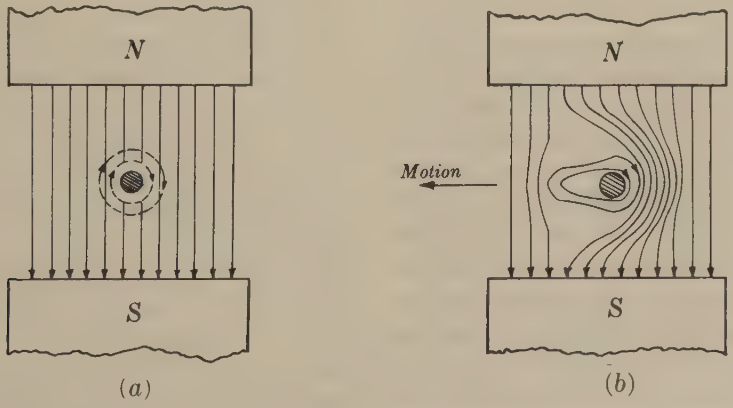


FIG.1-85 Electrode arrangement in an electromagnetic cathode-ray tube.

magnetic fields takes place. The two fields are shown separately in Fig. 1-86 (a) and the result of their interaction is shown in Fig. 1-86 (b). A force is exerted on the conductor which causes the conductor to move in the direction shown. This direction of motion is at right angles to the direction of electron flow in the conductor and to the direction of the field. However, if the current-carrying conductor is placed in the magnetic field with the conductor parallel to the lines of force of this field, the magnetic lines of force from the two sources are at right angles and do not interact. In this case no force is exerted on the conductor. If the current-carrying conductor is placed in the magnetic field so that the angle between its direction and that of the field is different from zero, the conductor experiences a force which increases as the angle increases.

The stream of electrons produced by the electron gun of a cathode-ray tube may be considered as equivalent to the stream of electrons in a solid conductor. Thus if the electron beam passes through a magnetic field the effect upon it is the same as upon the current-carrying conductor and the motion of the electron beam which results from interaction of the magnetic fields is at right angles to the direction of motion of the beam and to the direction of the field. This principle is applied in the magnetic focusing of the electron beam.

A focus coil is used in the electromagnetic cathode-ray tube to provide a



(a) (b)

FIG.1-86 Effect of placing a current-carrying conductor in a magnetic field.

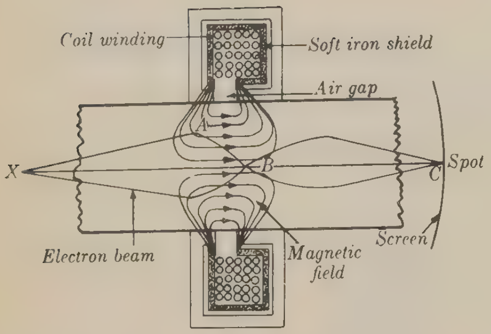


Fig. 1-87 Focusing action in an electromagnetic c.r.t.

field through which the electron beam passes. This focus coil is circular and is wound on a former which fits around the neck of the tube. Surrounding the coil is a soft-iron shield having an air gap on the inside next to the tube. The air gap allows the strong magnetic field to enter the tube, surrounding the electron beam with parallel lines of magnetic force. The soft-iron shield prevents this strong focusing field from interfering with the deflection fields. A direct current is passed through the coil to create the desired field. Some means of moving the focus coil along the neck of the tube is usually provided to permit approximate focusing adjustment. After this adjustment is made, usually in initial alignment, fine focusing is provided by adjustment of the amount of current flowing in the focus coil.

Fig. 1-87 shows the arrangement of the focus coil about the neck of the tube and the magnetic field set up inside the tube. Electrons of the beam proceeding from the cross-over point X along the line XC encounter the magnetic field. Since the field of these electrons is at right angles to the focus field no interaction takes place and the electrons proceed on to the screen without change in direction. However electrons proceeding from the cross-over point X along the line XA move diagonally through the magnetic field. The interaction of the magnetic fields causes the electrons to be deflected in a direction which is at right angles to their direction of motion and to the direction of the focus field. In this case for electrons along XAB and with the focus field as shown in Fig. 1-87 the direction of deflection is out from the page. However each electron still remains in a magnetic field and so it is continually deflected. The result is that the electron travels in a spiral path while it passes through the magnetic field. Electrons travelling on other diagonal paths across the field are also made to follow spiral paths. Electrons which are far off the axis are made to follow a path which causes them to meet at some point with the electrons that came into the field almost parallel with the axis. The position of this meeting point may be adjusted by varying the strength of the magnetic field so that it occurs at the fluorescent screen and optimum focus is obtained.

2. Deflection.

The effect of a magnetic field on the swiftly moving electrons of the electron beam has been reviewed. The magnetic field deflects the beam in a direction which is at right angles to the direction of the field and to the direction of the electron stream. As a result deflection of the beam may be obtained by providing two pairs of coils, one pair for horizontal and the other pair for vertical deflection. The pairs of coils are placed over the neck of the tube close to the focus coil (Fig. 1-85) and are sometimes called *deflection yokes*. The coils of a pair are mounted opposite each other and connected in series (Fig. 1-88). The magnetic field of each pair passes through the neck of the tube at right angles to the path of the electron beam. Thus the pair of coils arranged in the vertical plane provide horizontal deflection while the pair arranged in the horizontal plane

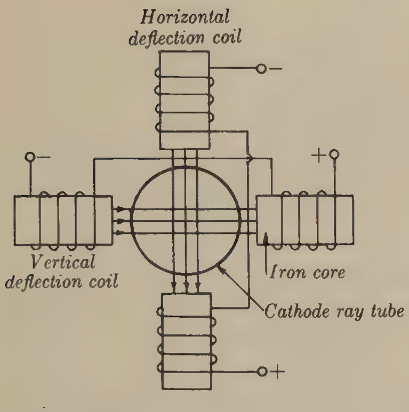


FIG.1-88 Deflection yokes of an electromagnetic c.r.t. for deflection in two directions.

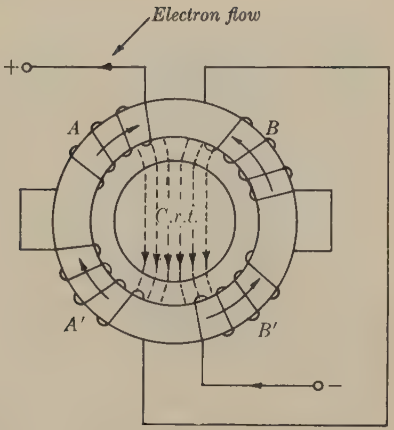


FIG.1-90 Circular deflection yoke.

provide vertical deflection. The electron beam, and so the spot on the screen, may be moved to any point by applying currents of the proper magnitude to these deflection coils. It is necessary in this deflection arrangement to provide good magnetic shielding between the focus coil and the deflection coils.

If deflection in one direction only is desired, coils may be wound on a single iron core placed about the neck of the tube (Fig. 1-89). The two coils are wound in series and in such directions that their magnetic fields in the iron core are in opposite directions. To complete the magnetic circuit, each field leaves the core and passes in the same direction through the neck of the tube. This uniform magnetic field passing across the tube, and so across the electron beam, provides the deflecting force on the beam. If the direction of motion of the electrons of the beam (Fig. 1-89) is out from the page the beam is deflected downward. If the current through the coils is increased, the field is increased and the beam is deflected farther downward. If the current through the coils is reversed the field is reversed and the beam is deflected upward. With this type of deflection coil deflection in any direction from the centre may be obtained by rotating the whole yoke around the neck of the tube.

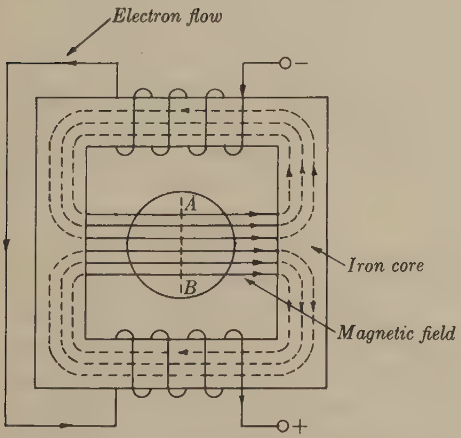


FIG.1-89 Deflection yokes for deflection in one direction only.

Fig. 1-90 shows a similar deflection arrangement which employs a circular yoke with two sets of identical coils symmetrically arranged on the core. For the position of the yoke and directions of field indicated the beam is deflected horizontally to the left.

For the electromagnetic cathode-ray tube the deflection sensitivity is expressed in millimeters per ampere, since deflection depends upon the magnetic field strength which depends upon the current in the coil. Thus the deflection

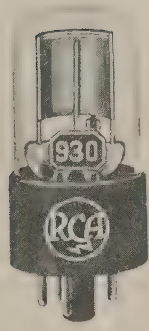


FIG. 1-91 A diode phototube.

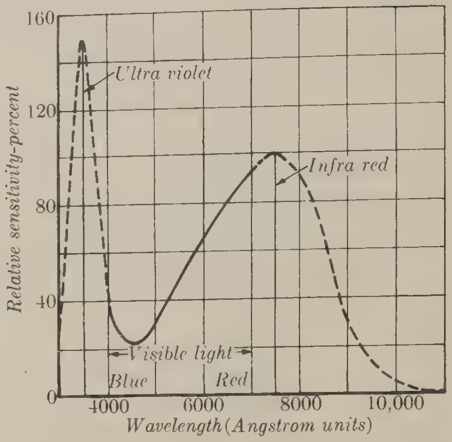


FIG. 1-92 Variation of sensitivity with the wavelength of the incident radiation for a phototube.

sensitivity is increased by increasing the number of turns of the coil. This results in an increase in inductance which is undesirable in high-frequency applications of such a tube.

1.29 The Phototube. The action of the *phototube* depends upon the phenomenon of photoelectric emission. Certain substances, for example, caesium oxide, emit electrons when exposed to radiation such as light, ultra-violet rays, infra-red rays, X-rays or radioactive rays. Fig. 1-91 shows a diode phototube. The curved cylindrical metal cathode has a caesium-oxide coating over its inner surface. When the radiation strikes the surface, electrons are emitted from it and are attracted towards the narrow anode rod, which is maintained at a positive potential with respect to the cathode. The number of electrons flowing from the cathode to the anode depends upon the intensity and the wavelength of the incident radiation, other factors being kept constant.

The sensitivity of a phototube is expressed as the number of microamperes of electron current per unit of light energy. For the valve pictured in Fig. 1-91, the variation of sensitivity with the wavelength of the incident radiation is shown in Fig. 1-92. The sensitivity is greatest for ultra-violet rays and quite high for infra-red rays. The sensitivities of phototubes must therefore be compared over the same small range of wavelength if the comparison is to be significant.

The sensitivity of a phototube is increased by replacing the high vacuum with gas at a low pressure. The electrons liberated from the cathode by the incidence of radiation travel towards the anode and if the anode potential is sufficiently large, the gas is ionized. The positive ions are attracted towards the cathode where their charge neutralizes the space charge around the cathode. Hence, when the gas is ionized, there is no limitation on the magnitude of the current except the number of electrons being emitted by the cathode. Therefore the anode current of a gas phototube is increased by as much as ten times that of a high-vacuum phototube for the same cathode surface, intensity and wavelength of radiation, and anode voltage. However, there is greater risk of accidental damage through positive-ion bombardment of the cathode at excessive anode voltages. Hence a gas phototube must be operated within more rigid temperature and voltage limits. Because of their greater sensitivity to changes in light intensity, gas phototubes are used

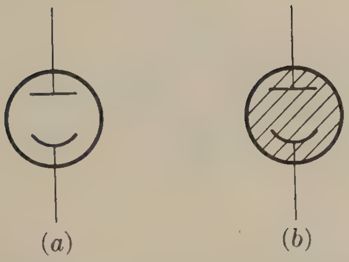


FIG.1-93 Circuit symbols for (a) a high-vacuum photo-diode (b) a gas photo-diode.

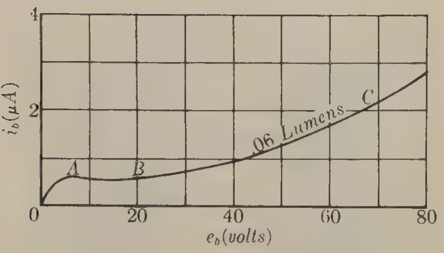


FIG.1-95 e_b-i_b curve for a gas phototube.

in a-f applications such as the sound track of a movie film. High-vacuum phototubes, on the other hand, are used in relay apparatus because of their greater reliability. Fig. 1-93 (a) and (b) shows the circuit symbols for a high-vacuum and a gas photo-diode respectively.

The anode voltage of all phototubes should not exceed about 15 to 20 volts. In a gas phototube, at higher anode voltages, positive-ion bombardment damages the cathode surface. In the vacuum phototube the slight amount of residual gas in the valve ionizes and the anode current fluctuates. Because the anode current is only of the order of microamperes these small fluctuations have a relatively large effect. Hence all phototubes must be connected in series with a voltage limiting resistor as in Fig. 1-94. Also because of the very small values of anode current, leakage currents become noticeable. Therefore leakage current such as that which exists between particles of dust on tube terminals, must be eliminated by careful cleaning of the terminals. The glass envelope is often coated with a suitable wax so that leakage currents caused by moisture films on the glass surface are avoided.

The e_b-i_b curve for a gas phototube is shown in Fig. 1-95 and that for a high-vacuum phototube in Fig. 1-96. The curve for the gas valve is not as regular as that for the high-vacuum valve. The anode current increases rapidly to A and no further increase occurs until the anode voltage reaches the ionization potential at point B. Beyond this point the anode current increases as the anode voltage increases. The anode resistance of the gas phototube, given by the reciprocal of the slope from B to C on its characteristic, is not as high as that of the high-vacuum phototube for the portion X to Y on its characteristic.

Fig. 1-97 (a) shows a gas phototube which is non-directional. Radiation from any direction penetrates the apertures and strikes the coated surface on the

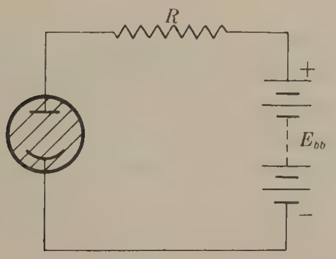


FIG.1-94 Circuit arrangement of a phototube.

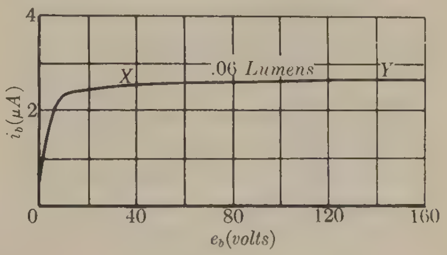
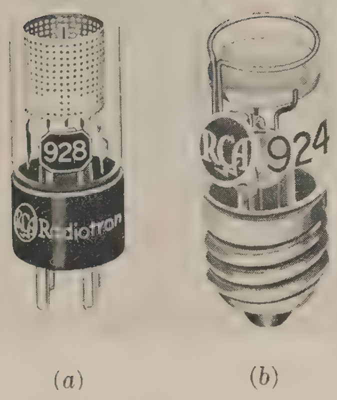


FIG.1-96 e_b-i_b curve for a high-vacuum phototube.



(a) (b)

Fig. 1-97 (a) Non-directional gas phototube (b) Gas phototube with a screw-in socket.

opposite side of the interior of the cathode. This gas phototube has a sensitivity of about 65 microamperes per lumen, a total cathode emission surface of 0.7 square inches, a maximum anode current of 15 microamperes with a maximum anode supply voltage of 90 volts. The series resistance must be from 1 to 4 megohms.

The anode consists of a circular ring to collect as many stray electrons as possible and the cathode is a metal hemisphere coated on the inside.

The sensitivities of gas phototubes range from 55 to 110 microamperes per lumen; those for high-vacuum phototubes range from 6.5 to 45 microamperes per lumen. Anode circuit supply voltages are usually 90 volts for gas valves but range up to 500 volts for high-vacuum valves. Anode currents range from 2 to 30 microamperes for both types. Another gas phototube with a screw-in socket is shown in Fig. 1-97 (b).

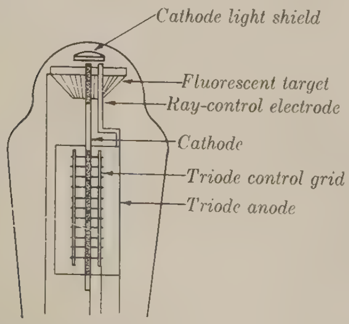


Fig. 1-98 Electrode arrangement in the tuning-indicator valve.

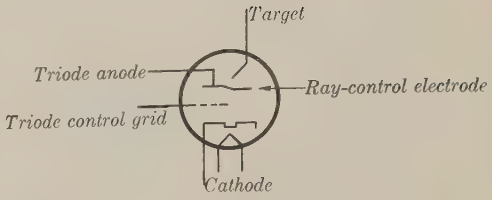


Fig. 1-99 Circuit symbol for a tuning-indicator valve.

1.30 The tuning-indicator valve. *The tuning-indicator valve* or *tuning eye* is usually employed to give a visual indication of accurate tuning of a receiver to an incoming signal. It consists of a cathode-ray indicator and a triode combined in a single valve envelope. Fig. 1-98 shows the arrangement of the electrodes and Fig. 1-99 the circuit symbol used for the combination. A single cathode projecting upward from the triode section into the indicator section serves for both units. The indicator section consists of the centrally located cathode surrounded by a funnel-shaped anode or *target* which is operated at a positive potential. The interior surface of the target is coated with a fluorescent material which glows a bright green when struck by electrons attracted from the cathode. The top end of the cathode is covered with a small round disc serving as a cathode light shield so that when the valve is viewed from above, the end of the valve appears as in Fig. 1-100 (a). The target is completely lighted and appears as a bright ring when there is a free passage of electrons from cathode to target.

To control the flow of electrons from the cathode to the target, a *ray-control electrode* (Fig. 1-98) consisting of a narrow metal cylinder is placed beside the

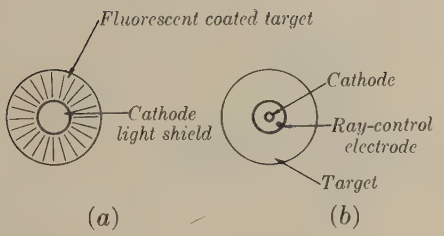


FIG. 1-100 Appearance of tuning-indicator valve when viewed from above.

cathode. When viewed from above, with the cathode shield removed, this electrode arrangement appears as in Fig. 1-100 (b). The ray-control electrode is operated at a positive potential less than, or the same as, the target potential. Varying the potential of the ray-control electrode varies the width of a shadow which it casts on the target.

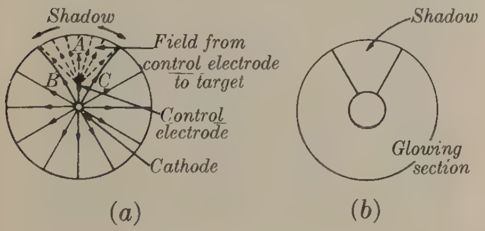


FIG. 1-101 Fields between electrodes in a tuning-indicator valve when the control electrode is considerably less positive than the target.

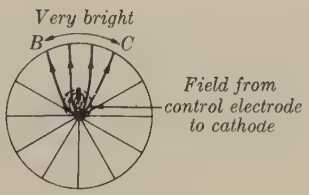


FIG. 1-102 Fields between electrodes in a tuning-indicator valve when control electrode is almost as positive as the target.

The explanation of the action is assisted by the drawing of some of the electrostatic lines of force which exist between the control electrode and the target. An electrostatic line of force, it is recalled, is the path along which a free electron would travel. The direction of a line of force is thus from a low potential to a high potential. When the control electrode is less positive than the target, there are many lines of force between these electrodes. These lines, the dotted lines in Fig. 1-100 (a), are practically radial straight lines since the target is circular. The electron paths from cathode to target are shown by solid lines. Electrons travelling directly toward the control electrode reach it. Electrons in nearby paths between *B* and *C* pass this electrode and tend to travel along the lines of force; that is, these electrons veer away from the control electrode toward the target as shown. No electrons reach the region of the target behind the control electrode. Since electrons from all other parts of the cathode travel along radial paths to the target, the remainder of the target glows. Hence a shadow is formed behind the control electrode. The tuning-indicator valve viewed from above in this condition appears as in Fig. 1-101 (b). If the control electrode is made considerably less positive than above, the number of lines of force between control electrode and target is increased causing the electrons to veer away from the control electrode to a greater extent. Thus the width of the shadow is increased. The angular width of the shadow may be increased to about 100° by making the potential of the control electrode decrease to almost that of the cathode. If the input control voltage is first amplified by a d-c amplifier, the potential of the control electrode may be made negative with respect to the cathode and a shadow width of about 180° can be obtained.

As the control-electrode potential becomes more positive, the shadow width decreases. When the control electrode is almost as positive as the target the field is concentrated between the control electrode and cathode as shown in Fig. 1-102 (a). There are practically no lines of force between the control electrode and the target. Hence more electrons are attracted towards the control electrode because of its high positive potential. This causes the electron paths to bend in towards the control electrode and more electrons strike a sq. cm. of the target between *B* and *C* than any other sq. cm. of the target so that this portion glows more brightly and the glowing wings appear to overlap.

The shadow width is exactly 0° and the brightness is uniform when the presence of the control electrode does not change the radial field which exists between the cathode and target. This condition occurs when the potential of the control electrode is equal to the potential at the position of the control electrode due to the potential gradient between cathode and target. Thus for zero shadow the control electrode has some intermediate value between anode and cathode potentials.

In practice, the valve is operated so that zero (or minimum) shadow width indicates accurate tuning. A circuit used for this purpose is shown in Fig. 1-103. The value of the controlling voltage which is applied between grid and cathode is determined by the tuning condition of the circuit. When the circuit is accurately tuned, this voltage has a maximum negative value. As a result, the anode current is reduced by this grid voltage to a small value. Hence the voltage drop across the resistor *R*, through which the anode current flows, is small. Since one end of the resistor is fixed at the potential of the positive supply voltage E_{bb}, the potential of the triode anode at the other end of the resistor is almost equal to E_{bb}. The ray-control electrode has the same potential as the triode anode since these two electrodes are connected internally. The target electrode is connected to the positive terminal of the supply voltage. Therefore with the controlling voltage a maximum, the ray-control electrode has a maximum potential. Hence the angular width of the target shadow is zero (or minimum) when the circuit is accurately tuned.

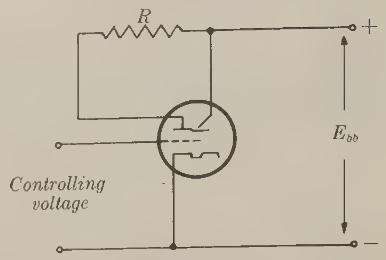


FIG. 1-103 Circuit arrangement of tuning-indicator valve.

When the external circuit is not accurately tuned, the controlling voltage is less negative. For this value of triode grid voltage, a larger anode current flows and a larger voltage drop appears across the resistor. This lowers the voltage on the control electrode, hence the shadow width is larger.

The change in the ray-control (and triode anode) voltage is much greater than the change in controlling voltage. This is possible since the triode section acts to amplify the controlling voltage before it is applied to the indicator section. For one type of double-unit valve electrode voltages and currents are as follows:

Heater voltage $6.3V$

Heater current $0.3A$

Target voltage $250V$

Maximum target current $4mA$

Resistance *R* 1 megohm

Max. triode anode current $0.24mA$

Triode grid voltage $-22V$ for 0° shadow angle; $0V$ for 90° shadow angle.

Ray-control electrode voltage $140V$ for 0° shadow angle; $0V$ for 90° shadow angle.

Some tuning-indicator valves are constructed to include only cathode, anode

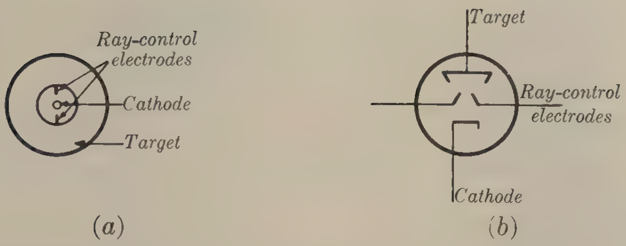


FIG. 1-104 Tuning-indicator valve which includes cathode, anode and two ray-control electrodes.

and ray-control electrode. Such a valve usually has two ray-control electrodes arranged on diametrically opposite sides of the cathode as in Fig. 1-104 (a). The two ray-control electrodes may be connected together to give two symmetrical but diametrically opposite shadows. An extra valve is required to amplify the controlling voltage before it is applied to this type of indicator valve. This valve may also be used to give two shadows, independentlv controlled by two different circuits. The circuit symbol for this type of valve is shown in Fig. 1-104 (b).

1.31 High-frequency valves. The valve designs already discussed are suitable for use at a-f, i-f and low r-f frequencies. However at higher r-f frequencies a change in valve design is necessary if a valve is to be able to produce a reasonable power output.

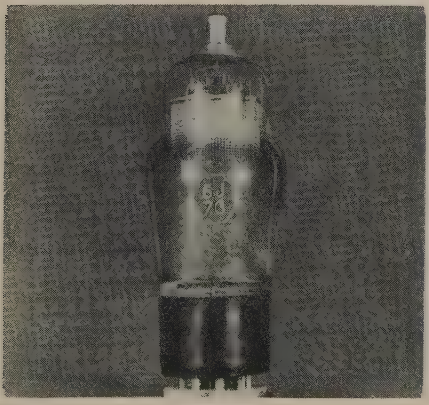


FIG. 1-105 Receiver pentode.

It has been pointed out that the triode construction described in section 1.10 is not useful at r-f frequencies because the grid-to-anode capacitance is so large that the valve is effectively by-passed. The tetrode and pentode valves were designed to minimize this capacitance and hence can be used at radio frequencies up to 1 or 2 Mc/s. In order to further reduce the grid-to-anode capacitance and the mutual inductance between leads, the grid and anode leads are separated as much as possible. By this construction, valves of the same size and having the same power-dissipating abilities can be used at frequencies up to 30 Mc/s. Fig. 1-105 shows a receiving pentode with the grid lead in the form of a short cylinder protruding through the top of the glass envelope. The anode lead extends through the glass seal in the tube base. Fig. 1-106 shows a transmitting triode which is approximately 9 inches high and has the anode lead emerging through the top of the envelope and the grid lead through the side. This type of grid and anode terminal construction is sufficiently rugged to withstand vibration. Mechanical rigidity of electrodes and leads is necessary since any changes in the relative spacings of electrodes or leads cause changes in the interelectrode capacitances and mutual inductances. Dielectric loss in valve bases is also serious at high frequencies. In high-power valves, such as that shown in Fig. 1-106, the anode and grid leads are sepa-

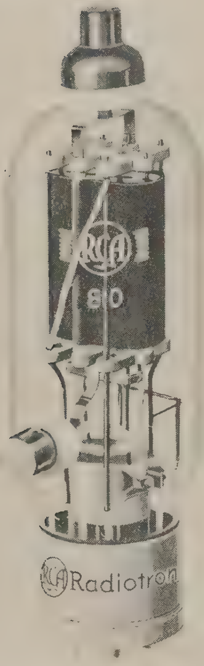


Fig.1-106 Transmitter triode (9 in. high).

rated for the additional reason that the anode is operated at a high positive potential.

Fig. 1-107 shows a triode which is designed to handle several thousand watts at frequencies up to 30 Mc/s. The valve employs post terminals which rigidly support the anode. With this type of terminal, no base is necessary and dielectric base power loss, which is serious at high frequencies, is eliminated. This valve is approximately $8\frac{3}{4}$ inches high.

At high frequencies power losses due to lengths of leads may be excessive. The use of very short leads in high-frequency valves reduces these power losses considerably. Interelectrode capacitance is reduced since there is a smaller electrode surface area. Shorter leads reduce the mutual inductance between electrodes. Electromagnetic radiation of r-f by the leads is also reduced. Fig. 1-108 shows a valve incorporating these features which can handle 30 or 40 watts at high frequencies up to 200 Mc/s. It is only $3\frac{1}{2}$ inches high. The power-handling capabilities of the valve shown in Fig. 1-107 have been sacrificed here for shorter leads and higher frequency-handling capabilities.

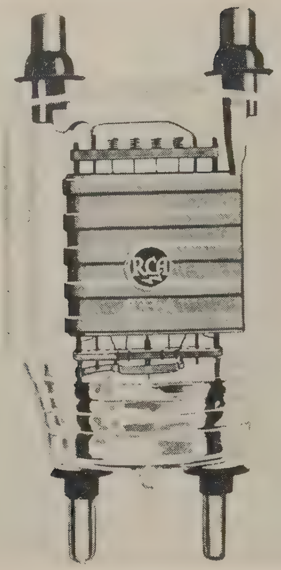


Fig.1-107 Transmitter triode ($8\frac{3}{4}$ in. high) used for frequencies up to 30 Mc/s.

At ultra-high frequencies the time taken for an electron to travel from cathode to anode in a valve must be taken

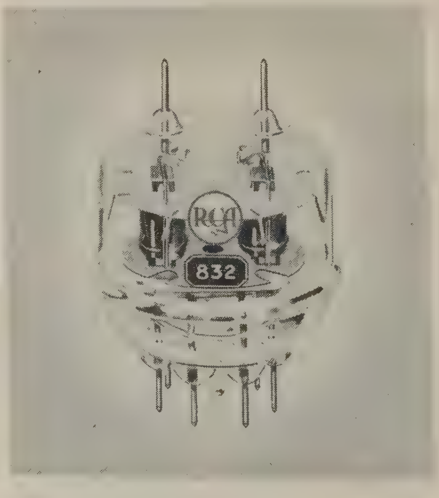


Fig.1-108 Transmitter valve ($3\frac{1}{2}$ in. high) used for frequencies up to 200 Mc/s.

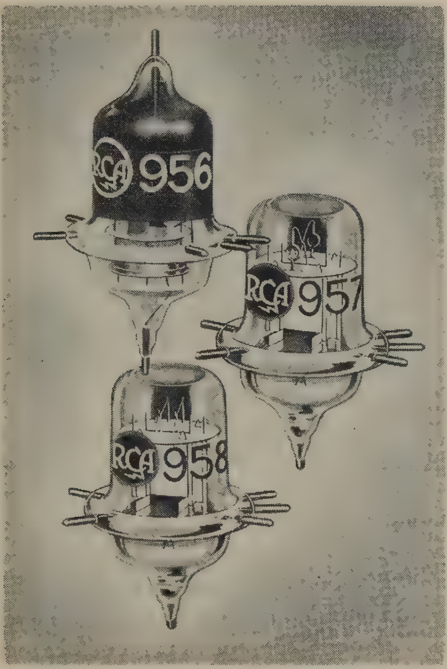


Fig.1-109 Acorn valves (1½ in. high) used for frequencies up to about 600 Mc/s.

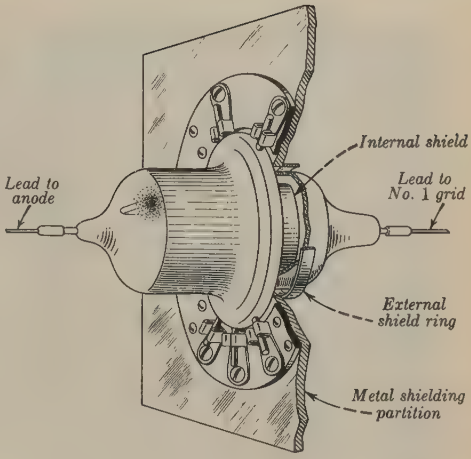


Fig.1-110 Method of connecting acorn valve in a circuit.

into consideration in valve design. This time interval is called the *electron transit time*. At very-high frequencies the transit time becomes appreciable in comparison to the period of an u-h-f alternating voltage. For example, at a frequency of 500 Mc/s, the time required for one cycle is .002 microseconds and for one half cycle is .001 microseconds. When the transit time is negligible, the anode current increases when the grid voltage becomes more positive. If the transit time is say, .001 microseconds, electrons that leave the cathode during the positive half cycle of grid voltage arrive at the anode during the negative half cycle of grid voltage and the increase in plate current is retarded one half cycle. Thus when the transit time is comparable to the period of the applied voltage, a phase shift between the grid voltage and anode current is introduced. This phase shift may seriously affect the normal operation of the valve. Electron transit time is decreased by decreasing the spacing between the anode and cathode. Fig. 1-109 shows a type of valve called an *acorn valve* having extremely small electrode spacing and short leads. The upper valve is a pentode less than two inches high while the two lower valves are triodes less than 1½ inches high. The triodes have amplification factors of about 12 at frequencies up to 600 Mc/s. Fig. 1-110 shows how an acorn valve is connected into a circuit by means of a supporting metal partition which also aids in separating the anode lead from the control grid lead. Fig. 1-111 shows an acorn thyratron held in a hand to show comparative size. Fig. 1-112 shows

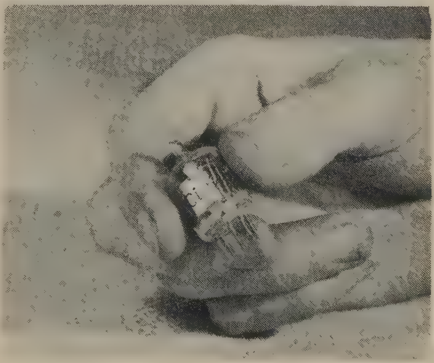


Fig.1-111 Acorn thyratron.

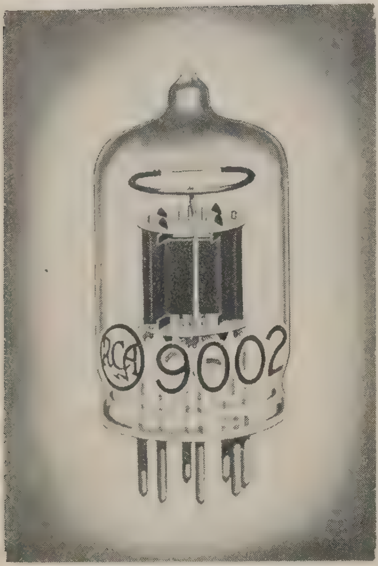


Fig. 1-112 Midget valve.

a *midget valve* which is similar to the acorn type. This triode has a higher amplification factor than an acorn triode and is also suitable for use at frequencies up to several hundred megacycles per second.

At frequencies of the order of thousands of megacycles per second, a different design of valve is necessary. Valves designed for these frequencies contain not only the valve electrodes but also some of the external circuit components in addition. Interelectrode capacitances and inductances are large enough at such frequencies that no additional inductors or capacitors are required. The *disk-seal valve*, the *magnetron* and the *klystron* are examples of this type of valve. The two latter valves are discussed in Chapter V.

The *disk-seal* valve or *megatron* is a high-vacuum valve employing a different electrode construction than the types previously mentioned. In most of the valves so far described, the electrodes have been arranged concentrically with the cathode in the center. In the disk-seal valve, however, the electrodes are flat, smooth-surfaced metal discs arranged in parallel planes. Fig. 1-113 illustrates the basic construction of a disk-seal triode. The heater is surrounded by a cylindrical cathode whose surface is increased by a metal cathode disc. As in the cathode-ray tube only the surface of the cathode which faces the anode is oxide-coated. The grid is a plane tungsten mesh parallel to the cathode emitting surface and is attached at the edges to a metal grid disc which is parallel to the cathode disc. The anode is a single, solid, metal cylinder extending from within the valve out through the glass seal. The interior and face of the anode is parallel to the grid and cathode and the spacing between cathode and anode is of the order of .005 inches. The anode, also, is attached to a metal anode disc. The glass walls are sealed to the electrode discs; this feature gives this type of valve its name. The silver-plated discs take the place of leads and have very little interlead inductance. All external metal portions of these valves are silver plated to provide low-resistance circuit connections, freedom from corrosion and minimum r-f losses because of eddy currents or skin effects. This construction makes a compact unit capable of withstanding severe mechanical

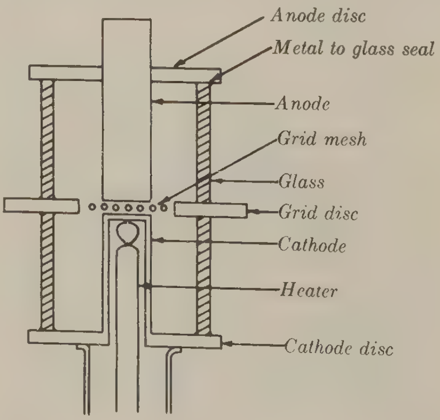


Fig. 1-113 Basic construction of a disk-seal triode.

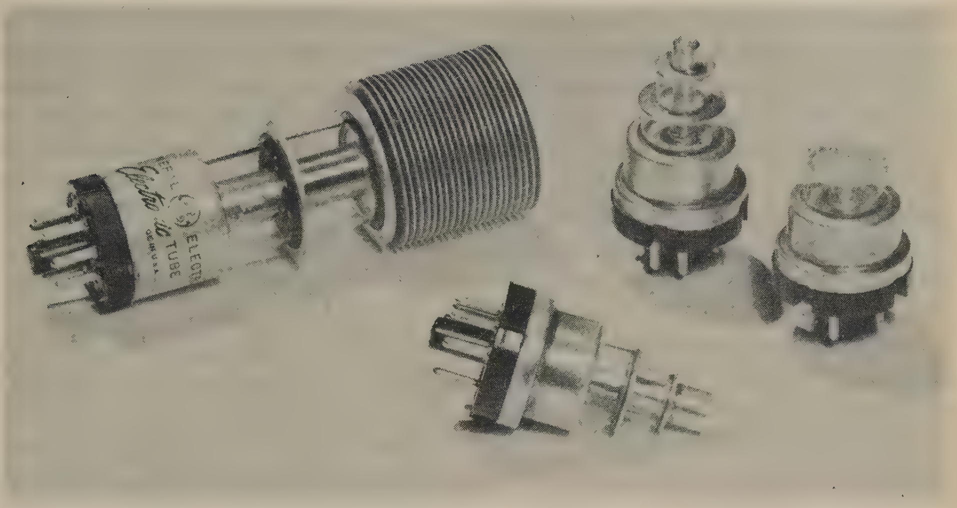


FIG.1-114 Four disk-seal valves.

stresses and providing reasonable power output at frequencies up to several thousand megacycles per second. Fig. 1-114 shows four disk-seal valves. The largest is a high-power valve having a finned anode for increased power dissipation. The three smaller valves called *lighthouse valves* are a special type of disk-seal valve in which the diameter of the electrode discs decreases away from the valve base. This pyramid structure is particularly adapted for the attachment of the output coupling devices used at these frequencies. Fig. 1-115 is a cutaway view of a lighthouse triode. In this valve the spacing between anode and cathode is about .007 inches.

A family of static $e_b - i_b$ curves for a lighthouse triode is given in Fig. 1-116.

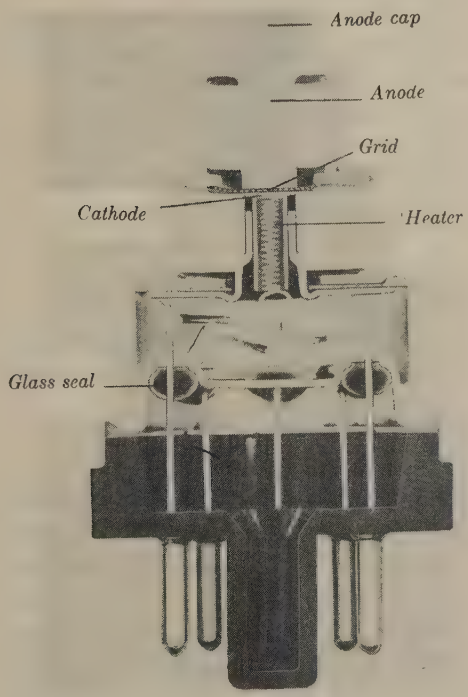


FIG.1-115 Cut-away view of a lighthouse triode.

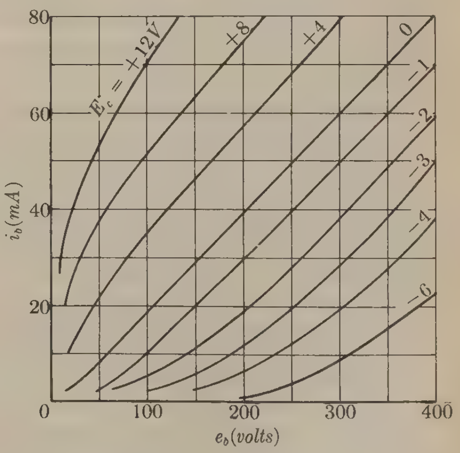


FIG.1-116 Lighthouse triode static $e_b - i_b$ curves.

Their shape is similar to the static $e_b - i_b$ curve studied in earlier sections. The curves shown are not extended to the anode saturation region since this type of triode is normally operated below saturation.

1.32 Valve nomenclature. During the initial period in the development of radio the naming of valves was a haphazard process. Eventually, however, a system for numbering and lettering valves was developed. In general, valves are classified according to their application. The three main classifications are:

(i) receiving valves
(ii) transmitting valves
(iii) other purpose valves.

A. Receiving Valves

The basic system used in designating receiving valves employs the following:

(i) A number to indicate the approximate filament or heater voltage, followed by
(ii) a letter to indicate the function of the valve, e.g. letters near the end of the alphabet indicate rectifiers, followed by
(iii) a number to indicate the number of elements of the valve connected to base pins and hence to the external circuit.

Along with the above the following suffixes are sometimes used to indicate other properties of the valve:

(i) A suffix $-G$ indicates a valve possessing a glass envelope and an octal base.
(ii) A suffix $-GT$ indicates a glass valve with an octal base having reduced dimensions.
(iii) The word metal appearing beneath a designation indicates the valve has a metal envelope and an octal base.
(iv) The letter S inserted after the number indicating heater voltage, but before the letter indicating the valve function, indicates a *single-ended valve*, in which all connections are made through the base of the valve. The valves thus designated have metal envelopes or glass envelopes with small dimensions.

However all valves are not designated by this system which is the commonest. In another system different valves are indicated by numbers only.

B. Transmitting Valves

There is no universal system of designation for transmitting valves. Standard types are designated by numbers of three or more digits. The higher numbers indicate more recently designed valves. These numbers however give no indication of the characteristics of the valves. Other series are designated by letter prefixes such as *RK* or *HY*

C. Other Purpose Valves

Under other purpose valves are classified cathode-ray tubes, television scan tubes, control and regulator valves.

1. CATHODE-RAY TUBES.

Cathode-ray tubes are usually designated to indicate the screen size and screen persistence. The first number in the valve designation gives the diameter of the screen in inches. The last letter and number indicate the persistence of the screen as follows:

*P*1 indicates medium persistence.
*P*2 indicates long persistence.
*P*3 indicates medium persistence especially suited for television.
*P*4 indicates the same as *P*3 but white in colour.
*P*5 indicates short persistence.

2. VOLTAGE-REGULATOR VALVES.

Voltage-regulator valves are usually designated by the letters *VR* followed by a number to indicate the operating anode voltage.

1.33 Valve ratings and data. Valve ratings and data pertaining to the operation of valves are discussed under the three main classes of valves.

A. Receiving Valves

The data usually given for a receiving valve is of three kinds: maximum ratings, typical operating conditions and electrical and physical characteristics.

1. Maximum Ratings.

These ratings are the maximum safe operating voltages, currents and powers for the electrodes. If these ratings are exceeded the life and performance of the valve are impaired. The ratings are as follows:

Heater voltage: the d-c value or the effective a-c value of the voltage applied to the heater terminals.

Heater Current: the d-c or effective a-c value of the current flowing through the filament at operating temperature and with the correct heater voltage.

Anode voltage: the d-c value of the voltage between anode and cathode. In the case of a diode rectifier, this is the effective value of the maximum a-c anode voltage.

Screen voltage: the d-c value of the voltage between screen grid and cathode.

Grid voltage: the minimum negative value of the d-c control-grid bias. For power-amplifier valves the peak value of the input signal is specified in volts.

Cathode current: in the case of multi-unit or multi-electrode valves the maximum total emission.

Anode dissipation: the maximum rate at which the anode is able to radiate energy or conduct it away safely as heat. This power rating is the difference between the d-c power supplied to the valve and the a-c power delivered by the valve to the load, in watts.

Screen dissipation: a power rating to indicate the rate at which energy is safely dissipated by the screen in the form of heat.

Peak inverse voltage: the peak value of the negative voltage which can be applied to the anode without an arc-back occurring.

Condensed-mercury temperature range: the maximum permissible temperature variation for a mercury-vapour diode rectifier. If this range is exceeded cathode disintegration may take place due to positive-ion bombardment.

Valve voltage drop: the maximum anode-to-cathode voltage in a gas valve. If this rating is exceeded, positive-ion bombardment of the cathode will occur.

2. Typical Operating Conditions.

A valve can be used under any suitable conditions within its maximum ratings. Where a valve may have different classes of operation, a set of typical operating conditions are given for each class. If the maximum ratings are not given separately they are indicated under the typical operating conditions. If no maximum ratings are shown the typical operating conditions are understood to be maximum ratings as well.

Under typical operation are listed electrode voltages and currents and the corresponding values of the anode resistance, trans-conductance and the amplification factor. For power amplifiers the load resistance and the power output are given. When cathode bias is specified the typical value of cathode resistor is given.

The following ratings sometimes occur under typical operation:

Anode current: the d-c value of the anode current.

Zero-signal d-c anode current: the d-c value of the anode current with no signal voltage applied to the control grid.

Maximum-signal d-c anode current: the d-c value of the anode current with a sine-wave signal applied to the control grid of sufficient amplitude to produce the rated voltage or power output.

Zero-signal and maximum-signal screen currents: the d-c value of the screen current with zero signal and maximum signal respectively applied to the control grid.

Total harmonic distortion: the percentage second harmonic distortion in the output waveform when the maximum-signal power output is being developed.

Maximum-signal power output: the product of the a-c components of anode voltage and current, less the power losses in the valve and load.

3. Electrical and Physical Characteristics.

(i) *Average anode characteristic curves.* The $e_b - i_b$ curves are drawn for several values of control-grid voltage. From these curves, the $e_c - i_b$ curves may be derived.

(ii) *Direct interelectrode capacitances.* These are the values of the grid-to-anode capacitance, the anode-to-cathode or output capacitance, and the grid-to-cathode or input capacitance. For multi-grid valves the values of the extra interelectrode capacitances are given. For multi-unit valves the values of the capacitances between the electrodes of one unit and those of the other unit are also given.

(iii) *Socket or base connections.* A schematic plan of the inverted valve base is given. The bottom view is shown since valve connections are usually made from below. For octal bases, the pins are numbered from 1 to 8 in a clockwise direction from the base key, as in Fig. 1-117 (a). For other types of bases the pins are numbered in the same direction beginning with the lower left hand filament pin which is numbered 1. In these bases the filament pins are usually thicker than the other pins, Fig. 1-117 (b), or are arranged as a pair with distinct spacing as in Fig. 1-117 (c).

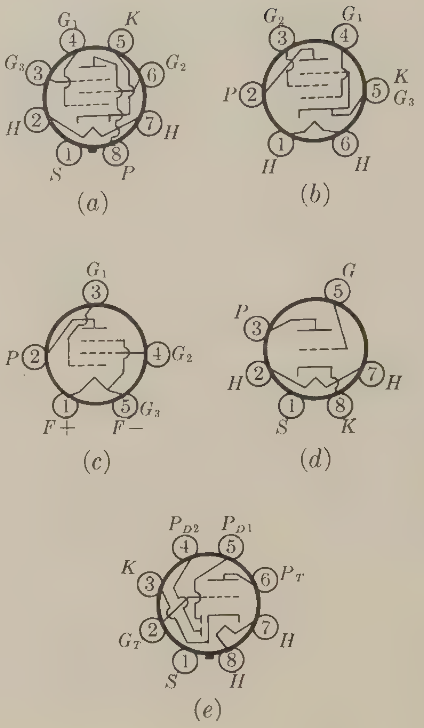


FIG. 1-117 Valve socket connections.

The octal base shown in Fig. 1-117 (d) is for a triode which has only five connected elements. Hence the other pins are left unnumbered. The letters H, K, G_1, G_2, G_3, P, and F indicate the connection to the heater, cathode, No. 1 grid, No. 2 grid, No. 3 grid, anode and filament respectively. In the case of a multi-unit valve the subscripts D, P, T and HX indicate whether the electrode belongs to the diode, pentode, triode or hexode units respectively as in Fig. 1-117 (e). The letter S indicates the pin for making an external connection to the metal shield in the case of a valve with a metal envelope. This shield is usually connected to ground so that stray charges which would otherwise collect on the envelope and affect the operation of the valve, are conducted to ground. The letters NC indicate that no connection is required.

B. *Transmitting Valves*

The information usually given for transmitting valves also consists of three kinds: maximum ratings, typical operating conditions and electrical and physical characteristics.

1. Maximum Ratings.

There are three kinds of maximum ratings: the *absolute maximum ratings*, the *maximum Continuous Commercial Service (C.C.S.) ratings* and the *maximum Intermittent Commercial and Amateur Service (I.C.A.S.) ratings.*

The absolute maximum ratings are the limiting values which, if exceeded, result in the shortening of the valve life and the impairment of its performance. The C.C.S. and the I.C.A.S. ratings are usually only given for air-cooled transmitting valves.

The C.C.S. ratings are the maximum values for valves to be used under continuous operating conditions. The I.C.A.S. ratings are the maximum values to be used where valves are operated for no more than 5 minutes continuously with stand-by periods of 5 minutes or more. Hence I.C.A.S. ratings are greater than C.C.S. ratings for the same total life.

The following ratings are usually given:

Filament voltage: the d-c or effective a-c value of the voltage applied across the filament. Variation in the filament voltage should not exceed ±5% unless otherwise specified.

Filament current: the d-c or effective a-c value of the filament current which flows at operating temperature.

Anode voltage, grid voltage, screen voltage, anode current, screen current, control-grid current: the d-c values at the given input voltage and power output.

Anode dissipation: the maximum rate at which the anode is able to radiate energy or conduct it away safely as heat.

Screen input: the product of the d-c screen voltage and the d-c screen current.

Anode input: the product of the d-c anode current and the d-c anode voltage.

Driving power: the average rate of energy dissipation in the grid and grid biasing circuit of the valve.

Peak a-f or r-f grid voltage: the peak value of the a-f or r-f signal voltage measured from the operating point.

Screen resistor: the value of the screen voltage dropping resistor required when the screen voltage is obtained from the anode power supply voltage. Its resistance is specified in place of the screen voltage.

2. Typical Operating Conditions.

The d-c values of the operating voltages and currents for each electrode are given. The driving power and the output power corresponding to these conditions are also given. When no other maximum voltages are given, the typical operating conditions are the maximum conditions. These conditions hold for only one class of operation. A different set of operating conditions is given for each class of operation or application.

3. Electrical and Physical Characteristics.

(i) *Direct interelectrode capacitances.* The meanings of these values are the same as that for the capacitances given in the case of receiving valves.

(ii) *Type of base and socket connections.* A diagram is drawn of these connections and the type of base is indicated in the diagram.

Finally a short description of the functions or possible applications of the valve are given along with the dimensions of the valve.

1.34 The valve tester. The best method of checking the performance of a valve is to compare its performance with that of a new valve under the same operating conditions. The performance test is the only complete test. However an auxiliary valve tester is of practical use when equipment is completely inoperative. The valve tester should be capable of making one or more of the following tests or checks:

(i) a check to indicate whether a short circuit exists between any pair of electrodes
(ii) a transconductance test
(iii) a check of the anode and screen currents under typical operating conditions
(iv) a test of the power input.

The *short-circuit test* should be made first since the application of operating electrode voltages to a valve containing a short may damage the test apparatus. The *transconductance test* is more indicative of the condition of a valve than an emission test. The emission of a valve may vary widely but may not seriously affect the operation of the valve. Also faulty valves often show full emission. The transconductance test takes into account one of the operating constants and hence gives a fairly accurate idea of the actual operating condition of a valve.

Most valve testers are not constructed to make all the above tests. Circuits capable of making each of the above tests are discussed in the following sections.

A. The Short-Circuit Test

A circuit which tests for shorts between electrodes is shown in Fig. 1-118. The input a-c voltage is stepped down by means of a step-down transformer T with a low-voltage secondary. By means of the switch S the secondary voltage may be applied across different pairs of electrodes. If two electrodes are shorted the lamp is lighted by the flow of current through the short. The lamp used should be one which operates on a very small current since the electrodes may be sufficiently close to give a high-resistance connection. The valve should be heated by the flow of heater current since the short may occur only when the valve and electrodes are hot. The potentiometer R varies the secondary filament voltage in order that the filament voltage may be maintained at its correct operating value as indicated by voltmeter V. A tetrode valve is shown but a similar circuit may be designed for other valves.

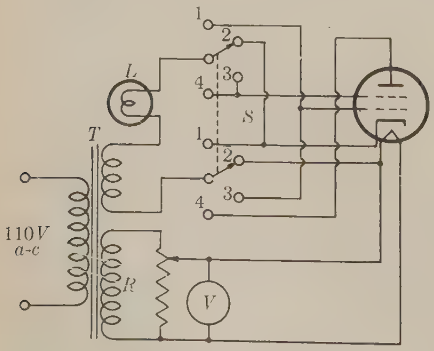


FIG.1-118 Circuit for conducting short-circuit test between valve electrodes.

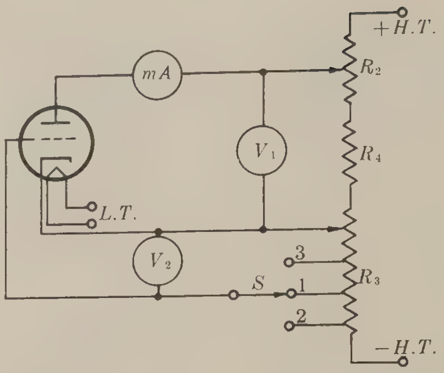


FIG.1-119 Circuit for measuring the static transconductance of a triode.

B. The Transconductance Test

Fig. 1-119 shows a circuit suitable for measuring the static transconductance of a triode. The operating voltages for the valve are obtained by the setting of three potentiometers. The filament voltage is obtained in a similar manner as in Fig. 1-118 and is therefore not shown. The tap on R_2 sets the anode voltage at a typical operating value as read on V_1. The control grid bias is set by R_3 with S in position 1 as read on V_2. The valve is now operating and the anode current is measured by the milliammeter (mA).

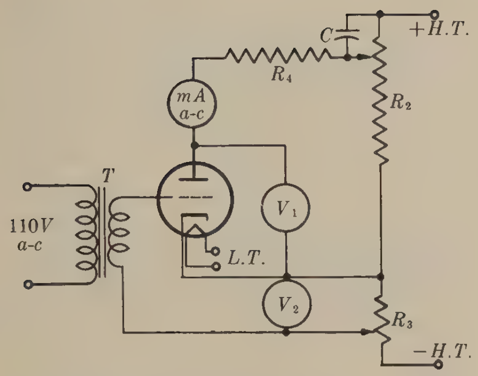


FIG. 1-120 Circuit for measuring the dynamic transconductance of a triode.

The transconductance of the valve is obtained by changing the grid voltage and observing the corresponding change in anode current. This is done by moving *S* from position 2 to position 3. In position 2 the grid bias is increased from position 1 and the decreased anode current is noted. In position 3 the grid bias is decreased from position 1 and the increased anode current is noted. The static value of the transconductance in mhos is then given by

$$g_m = \frac{\text{difference in the anode current readings (in amperes)}}{\text{difference in the grid voltage values (in volts)}}$$

Positions 2 and 3 of the switch may be preset so that the grid voltage change is exactly 1 volt. Then the value of the transconductance in mhos is equal to the change in anode current in amperes.

A more useful test is one which checks the dynamic transconductance since it is the dynamic constant which is important in actual operation. Fig. 1-120 shows a circuit for measuring the dynamic transconductance of a triode. As before the anode voltage and grid voltage are set by R_2 and R_3 and read on V_1 and V_2. However a load resistance R_4 is placed in the anode circuit and an a-c signal voltage of 1 *V* is applied to the grid via the secondary of an a-c transformer *T*. The varying grid voltage causes the anode current to vary. The a-c milliammeter *mA* reads the value of the a-c component of the anode current flowing through the load R_4. The dynamic transconductance value in mhos is then given by

$$g_m = \frac{\text{the a-c component of the anode current (in amperes)}}{\text{the signal voltage (in volts)}}$$

$$= \frac{\text{a-c component of anode current}}{1}$$

$$= \text{a-c component of anode current.}$$

C. Anode and Screen Current Check

In either of the previous two circuits of Figs. 1-119 and 1-120, the operating value of the current can be checked. With the anode voltage and the grid bias set at typical operating values by means of the potentiometers R_2 and R_3 respectively, the value of anode current may be noted on the anode milliammeter *mA* and compared with the standard value from a table of operating values. In the case of a tetrode or pentode, the screen current can be checked in a similar manner using the circuit shown in Fig. 1-121. The tap on R_4 sets the screen voltage at its correct operating value and the milliammeter mA_1 indicates the screen current value

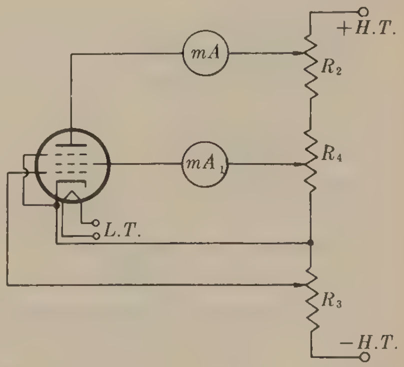


FIG. 1-121 Circuit for checking anode and screen current of a valve.

which can be checked by comparison with a table of normal operating values.

D. The Power Output Check

The power output test gives the best indication of the performance for a power valve. Fig. 1-122 shows a circuit suitable for measuring power output of a valve which is operated so that anode current flows all the time. The heater voltage is set at its correct operating value. The taps on R_2 and R_3 set the anode voltage and the grid bias respectively at their correct operating values. The potentiometer R_6 varies the a-c input voltage so that the grid voltage never swings beyond cut-off. The a-c component of anode current is made to flow through the load resistance R_5 by the high reactance of the choke L. The capacitance C offers little reactance to the a-c component of anode voltage but blocks the d-c component. The value of the a-c component of anode current is read by means of the a-c milliammeter mA. From the current value and the known value of the load resistance R_5, the value of the a-c power output (I^2R) developed in the load resistor can be calculated. Similar circuits may be designed to measure the power output for other types of valves.

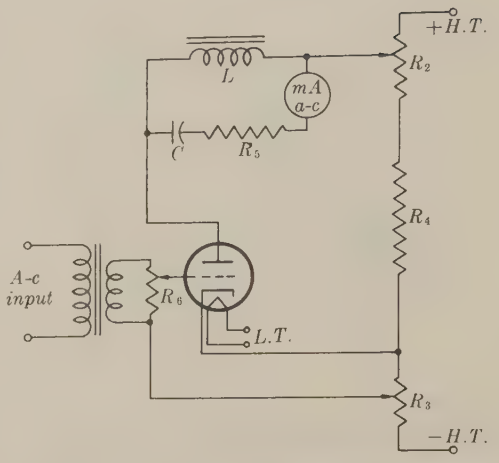


FIG.1-122 Circuit for measuring power output of a valve.

1.35 Questions and problems.

1. Define the terms: cathode, anode and electrode.
2. Name four ways in which electrons are emitted from a substance. Describe each briefly.
3. State three reasons why a valve must employ either gas at very low pressure or a high vacuum.
4. (a) What is meant by positive-ion bombardment?
   (b) Why is it an undesirable effect?
5. What is meant by the terms
   (a) hard valve
   (b) soft valve?
6. Explain the effect known as occlusion of gases and state why it is an undesirable effect in valves.
7. Draw diagrams to illustrate the difference between a directly-heated cathode and an indirectly-heated cathode. Give the advantages and disadvantages of each type.
8. Explain the term emission efficiency.
9. Name the three types of cathode materials. Compare them with respect to emission efficiency and durability and hence discuss the restrictions on their applications.
10. A tungsten cathode has an area of $\frac{1}{2}$ sq. cm. It is connected to a $10V$ d-c source and the filament current which flows is $0.5A$. If the emission efficiency is $10mA$ per sq. cm. per watt calculate the electron emission current in amperes.
11. An oxide-coated cathode is heated by the passage of a direct current of $0.3A$ through a filament whose hot resistance is 21 ohms. The emission efficiency is $75mA$ per sq. cm. per watt. The cathode is cylindrical with a diameter of 1 mm. and a length of 1 inch. Calculate the electron emission current from this cathode in mA.

12. (a) Define the term anode dissipation.
    (b) How may the anode dissipation of a valve be increased?
13. Explain why the anode in a high-power valve must be cooled. State two types of cooling.
14. State five desirable characteristics which an anode material should possess.
15. Name three commonly used anode materials and compare them with respect to the characteristics mentioned in question 14.
16. Describe the two processes used in the complete evacuation of a valve envelope.
17. Explain why induction heating cannot be used with metal valves to assist in the evacuation process. What type of heating is used with metal valves?
18. Explain why a getter may not be used with high-power valves.
19. Draw the circuit symbols for
    (a) directly-heated diodes
    (b) indirectly-heated diodes.
20. Draw a diode with a directly-heated cathode in a circuit including the battery supplies for L.T. and H.T. and a resistor in series in the anode circuit. Indicate the directions of the electron flow in these circuits.
21. Define the term valve characteristic.
22. (a) Draw the circuit which may be used to obtain the characteristic curve of a diode. Explain the use of each circuit component.
    (b) Explain the procedure used in obtaining the characteristic curve in (a).
23. Make a diagram to show the shape of the anode characteristic of a diode. Explain its shape.
24. What is the space charge around a heated cathode? Explain its effect on the anode current as the anode voltage is increased.
25. Define the terms: anode saturation, anode voltage, diode resistance.
26. Make a sketch to illustrate the construction of a triode.
27. Describe the construction of the control grid in a triode.
28. Describe the action of the control grid when the grid voltage is varied from a positive value to cut-off for a steady value of anode voltage.
29. Draw a circuit by means of which the static $e_b - i_b$ curves of a triode may be obtained.
30. Make a diagram to show the shape of the static $e_b - i_b$ curves. State why more than one curve is usually drawn.
31. Explain how the static $e_c - i_b$ curves of a triode may be obtained from the circuit in question 29. Explain their shape briefly.
32. (a) Using the values in the following table, plot the five static $e_b - i_b$ curves.

$E_c = 0V$

e_b	i_b
$40V$	$1mA$
65	2
86	3
112	4
133	5
155	6

$E_c = -1V$

e_b	i_b
$70V$	$1mA$
100	2
125	3
150	4
170	5
190	6

$E_c = -2V$

e_b	i_b
$110V$	$1mA$
140	2
165	3
185	4
210	5

$E_c = -3V$

e_b	i_b
$140V$	$1mA$
170	2
200	3
220	4
240	5

$E_c = -4V$

e_b	i_b
$170V$	$1mA$
210	2
230	3
260	4

(b) Using these curves draw a graph showing three static $e_c - i_b$ curves for $E_b = 100V$, $150V$ and $200V$ respectively.

33. Define the three triode factors.

34. Explain how two of these factors are obtained from the linear portions of the characteristics.

35. Explain how g_m and r_a are obtained at some point on the curved portion of the characteristics.

36. Explain the significance of μ.

37. State typical values for each factor for a triode.

38. (a) From the curves drawn in question 32(a) calculate the value of r_a on the linear portion for $E_c = -1V$ and $E_b = 150V$.
    (b) From the curves drawn in question 32 (b) calculate the value of g_m on the linear portion for $E_b = 150V$ and $E_c = -1V$.

39. (a) State the relationship between the three factors of a valve.
    (b) Calculate the value of μ using the values of g_m and r_a obtained in question 38.

40. Calculate the values of the three triode factors from the characteristics obtained in question 32 for the following operating voltages: $E_c = -3V$ and $E_b = 200\ V$.

41. The transconductance of a triode is 1600 micromhos, and the amplification factor is 20. Calculate the change in anode current resulting from a change in anode voltage of 50 volts. The values of the factors and any changes in voltage are for the linear portions of the characteristics.

42. Explain with the aid of a diagram the presence of three inter-electrode capacitances in a triode. Give typical values for each.

43. Explain the purpose, position and action of a screen grid in a tetrode.

44. Draw a circuit which may be used to study the static characteristics of a tetrode.

45. Sketch and explain the shape of the static $e_b - i_b$ curve of a tetrode.

46. (a) Using Fig. 1-45, calculate the value of the negative anode resistance of the tetrode (over the linear portion).
    (b) If the transconductance is 1700 micromhos calculate the amplification factor using the value of r_a found in (a).

47. State over which portion of the anode characteristic a tetrode is usually operated and explain why.

48. Explain why the anode resistance of a tetrode under normal operating conditions is much higher than that of a triode.

49. Make a sketch to show the components of a pentode.

50. (a) Explain the position, purpose and action of the suppressor grid.
    (b) State the source of operating voltage for this grid.

51. (a) Sketch a static $e_b - i_b$ curve for a pentode.
    (b) Explain why it differs from that of a tetrode.

52. (a) Explain over what portion of the $e_b - i_b$ curve a pentode is usually operated.
    (b) Give a typical value of the anode resistance of a pentode.

53. The first of the following tables gives corresponding values of grid voltage and anode current for constant anode and screen voltages. The second gives corresponding values of anode current and anode voltage for constant values of screen and control-grid voltages.

$E_b = 200V$ $E_{c2} = 100V$

i_b in mA	11.5	10.1	8.4	7.0	5.7	4.6	3.6	3.0	2.3	1.8	1.1	0.8
e_c in V	0	−1	−2	−3	−4	−5	−6	−7	−8	−10	−15	−20

$E_c = -3V$ $E_{c2} = 100V$

i_b in mA	0.2	5.5	6.3	6.7	6.8	6.9	7.0	7.1	7.15
e_b in V	0	20	40	80	120	160	200	240	280

(a) Plot the values given above and hence obtain a static $e_b - i_b$ and a static $e_c - i_b$ curve.

(b) Calculate the transconductance, anode resistance and amplification factor for $E_c = -3V$, $E_b = 200V$, and $E_{c2} = 100V$.

(c) Repeat these calculations for the same anode and screen voltages but for $E_c = -10V$.

(d) State to what type of valve the above characteristics likely belong.

54. (a) Explain the position, purpose and action of the beam-forming plates in a beam power valve.
(b) Explain why the anode current in a beam power valve rises to its full value for a lower anode voltage than in a standard pentode. State the advantage of this characteristic.

55. Describe the construction of the control grid of a variable-mu pentode and explain how this construction affects the anode current with varying grid voltage as compared with an ordinary pentode.

56. Explain how the amplification factor changes with the grid bias changing from zero to cut-off in a variable-mu pentode.

57. (a) Explain the difference between the terms multi-unit valve and multi-electrode valve.
(b) Draw circuit symbols to show two types of each of these two groups of valves.

58. State three effects in a valve due to the presence of gas.

59. Explain why a very small anode voltage in a gas valve causes a saturation current to flow.

60. Explain the term ionization potential.

61. Explain the purpose of the series resistor used with gas-filled valves.

62. The ionization potential of a gas diode is $21V$ and the saturation current is $4A$. Calculate the smallest value of load resistor to be connected in series with the diode when the supply voltage is $150V$.

63. Explain why the filament of a large gas valve must be switched on before the anode voltage is switched on.

64. Explain the purpose of the heat-shield cylinder often placed about the cathode in a gas valve.

65. (a) What causes arc-back in a gas diode?
(b) How does the maximum peak inverse voltage of a gas diode change as the temperature increases?

66. Explain why current flows in a cold-cathode gas valve.

67. What is the meaning of the term breakdown voltage in a cold-cathode gas diode?

68. Draw circuit symbols of
(a) hot-cathode gas diode
(b) cold-cathode gas diode.

69. (a) Explain the condition in a thyratron as the anode voltage is increased from zero to the ionization potential.

(b) Why has the control-grid voltage no further control over the anode current once saturation current flows?
(c) How may the anode current be reduced to zero?

70. (a) Sketch one type of construction of a thyratron.
(b) Draw its circuit symbol.

71. Compare the anode resistance of a thyratron before ionization with its value after ionization.

72. Make a diagram to show the essential components of an electrostatic type of cathode-ray tube.

73. Describe the construction and function of the following parts of a cathode-ray tube:
(a) cathode
(b) control grid
(c) accelerating electrode
(d) fluorescent screen
(e) envelope.

74. Describe the construction, function and action of the anodes in an electrostatic c.r.t.

75. Describe the position and action of the deflection plates in an electrostatic c.r.t.

76. Show the position of the spot on the screen of a c.r.t. when equal positive voltages are applied to the upper Y-plate and to the left X-plate.

77. What is observed on the screen if the frequency of an a-c voltage applied to the X-plates is increased from 2 c/s to 30 c/s?

78. Define sensitivity as applied to an electrostatic c.r.t. and give the unit of measurement used.

79. Give two functions of the aquadag coating on a c.r.t.

80. Explain why the second anode is usually operated at ground potential.

81. Describe the construction and action of the focus coil in an electromagnetic c.r.t.

82. Describe the construction and action of the deflection coils when
(a) there is one pair of rotatable coils
(b) there are two pairs of stationary coils in an electromagnetic c.r.t.

83. Explain how a phototube conducts an electric current.

84. What is one difference in operation between the high-vacuum phototube and the gas phototube?

85. Draw a diagram of a tuning indicator valve and explain its action
(a) in producing a shadow
(b) as the shadow width increases.

86. How does a tuning indicator valve indicate accurate tuning?

87. Draw circuit symbols for a gas phototube, a beam power pentode, a variable-mu pentode, a tuning indicator valve.

88. Describe what changes in the design of valves are necessary in order to make them suitable for use at very-high frequencies.

89. Why must the electrodes and leads of an h-f valve be very rigid mechanically?

90. Explain the term electron transit time and why its effect is undesirable at ultra-high frequencies.

91. Explain briefly how the disk-seal valves differ in construction from the usual valve.

92. Explain the significance of the numerals and letters in the following valve designations: 12SQ7-GT; 117Z6-GT; 5Z3; 6J8-G; 6L6 *metal*; VR150; 9AP4.

93. Explain the difference between absolute, C.C.S. and I.C.A.S. ratings for transmitting valves.

94. Explain the significance of the following ratings: anode dissipation, maximum signal d-c anode current, peak inverse voltage, maximum signal power output, anode input, driving power.

95. Explain the markings on the diagram of socket connections in Fig. 1-117 (a), (b) and (c).

96. Which test should be made first in checking a valve and why?

97. Explain why a dynamic transconductance test is preferable to a static transconductance test.

98. Explain why the power output test gives an excellent indication of the condition of a valve.

99. Draw a circuit suitable for
    (a) a short-circuit test
    (b) transconductance test
    (c) power output test.

Letter Abbreviations and Symbols
(Valve Applications)

QUANTITY	SYMBOL	SUBSCRIPT	EXAMPLE
D-c components	E I	b anode c control grid	E_b, I_b E_c, I_c
Direct supply voltages	E I	b_b anode c_c control grid	E_{bb} E_{cc}
Effective values of alternating components	E I	a anode g control grid g_2 screen grid	E_a I_a E_g I_g
Peak valves of alternating components	E_{max} I_{max}	a anode g control grid g_2 screen grid	$E_{a\ max}$ $I_{a\ max}$ $E_{g\ max}$ $I_{g\ max}$
Instantaneous values of alternating components	e i	a anode g control grid g_2 screen grid	e_a i_a e_g i_g
Instantaneous values of the total voltage or current (the sum or difference of inst. d-c and a-c components).	e i	b anode c control grid	e_b i_b e_c i_c

CHAPTER II

AMPLIFICATION

The principle of amplification is utilized in almost every type of electronic equipment. For example audio amplifiers make it possible to have one's voice heard with sufficient volume to be intelligible by large audiences indoors or outdoors. Radio-frequency amplifiers make it possible to increase the power of tiny oscillations from low-power valve circuits to the level at which they may be propagated over thousands of miles. Even a simple radio receiver is capable of producing amplification of the order of one million times. In all these devices the amplification principle is employed. Amplification is made possible by the use of the vacuum tube or valve, which forms the heart of any amplifier circuit.

2.1 Operation of a simple amplifier. A simple amplifier employing a triode valve is shown in Fig. 2-1. It is characteristic of amplifiers used to amplify voltages in the audio-frequency range. The three sources of supply voltage shown as batteries may also be generators or electronic power supplies which provide the voltages indicated. The heater or filament which is used to heat the cathode of the valve is connected to the *L.T.* or *"A" supply*. The anode of the valve is connected through a resistor *R*, called the *anode load resistor* to the positive terminal of the *H.T.* or *"B" supply*. A battery is connected between the grid and the cathode to make the grid negative with respect to the cathode. This battery is called the *bias battery* or *"C" supply* and the voltage is referred to as the *grid bias*. The voltage to be amplified is an a-c voltage which is applied in series with the bias between the grid and the cathode of the valve. This voltage is called the *grid signal* or *input signal* and is represented in the figure by a generator. The amplitude of the grid signal is normally less than the bias for this type of amplifier.

The *output voltage* of the amplifier is the a-c voltage produced across the load resistor *R* as the result of the a-c signal voltage applied to the grid. The operation of the circuit during one cycle of sine wave input voltage is as follows. When the amplitude of the signal is zero at the beginning of the cycle, the grid potential is 10 volts negative to the cathode (Fig. 2-1). The electron current from the *B* battery flows from cathode to anode through the valve and also through the load resistor. This current develops a p.d. across *R* so that part of the battery voltage appears across *R* and the remainder between cathode and anode of the valve. When the signal voltage increases during the positive half cycle, the grid potential becomes less than 10 volts negative to the cathode.

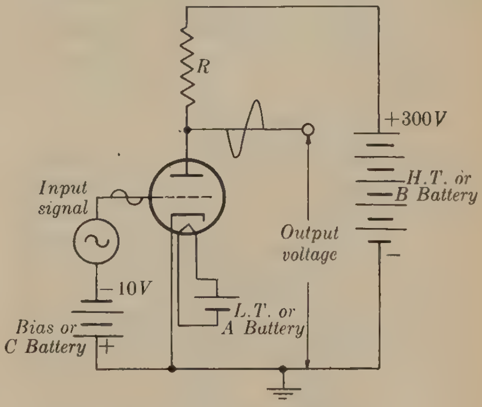


FIG.2-1 A simple amplifier.

Current increases through the valve and the p.d. across R increases. Since the potential at the upper end of R is maintained at $300V$ by the battery, an increase in p.d. across R results in a decrease in p.d. across the valve. Thus, an increase in the signal voltage causes the anode potential to fall. When the grid signal returns to zero at the end of the positive half cycle, the current through the valve and hence the anode potential return to their original values. When the signal voltage becomes negative during the second half of the input cycle, the grid potential becomes more than 10 volts negative to the cathode. The current through the valve decreases and so the p.d. across R becomes less. Hence the anode potential rises. Finally, as the signal returns to zero at the end of the input cycle, the current through the valve increases to its original value and the anode potential falls to its original value. The complete cycle of the grid voltage changes gives rise to a corresponding cycle of anode voltage changes. The amplitude of voltage variations at the anode are normally larger than the voltage variations at the grid. Consequently the circuit operates as a *voltage amplifier.*

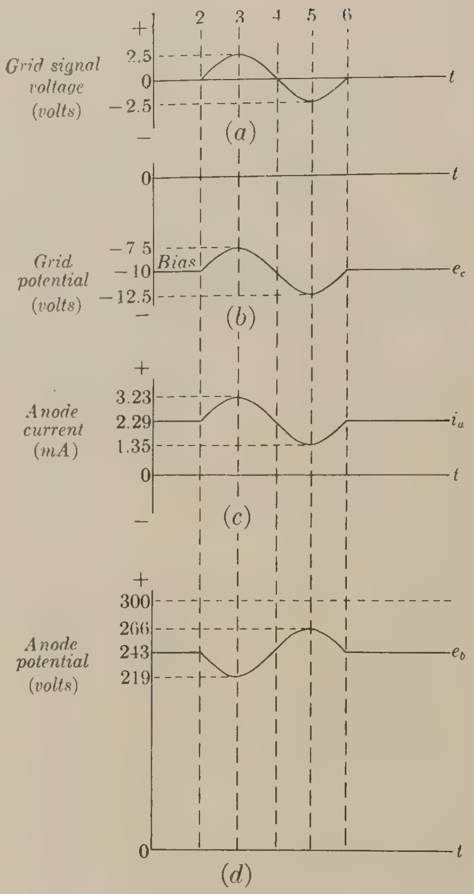


FIG. 2-2 Waveforms of voltage and current in an amplifier.

A more continuous picture of the operation of the amplifier is provided by the waveforms of Fig. 2-2. During the period from 1 to 2 the valve operates with steady values of grid voltage, anode current and anode voltage. The grid-signal voltage remains constant at zero. This condition is represented by the straight horizontal line through the zero voltage point in graph (a). The grid bias also remains steady at −10 volts as indicated by the horizontal line from 1 to 2 through $-10V$ in graph (b). The steady grid voltage results in a steady flow of anode current. This condition is represented by the horizontal line through $2.29mA$ in graph (c). A steady p.d. of 57 volts is developed across the anode load resistor which maintains the anode potential at 243 volts. This is represented by the horizontal portion of graph (d).

The first quarter of the grid voltage cycle is represented by the region from 2 to 3 on each graph. During this time the signal voltage rises from zero to 2.5 volts as indicated in graph (a). The grid potential therefore rises from $-10V$ to $-7.5V$ as shown in graph (b). This causes the anode current to increase from $2.29mA$ to $3.23mA$, graph (c). The p.d. across the anode load resistor increases causing the anode potential to fall from 243 to 219 volts as shown in graph (d).

The second quarter of the grid cycle is shown in the region from 3 to 4. During this time the grid signal returns to zero and the grid potential returns to the steady bias value of −10 volts. This condition is shown at position 4 in graphs (a) and (b). The decrease in grid voltage causes the anode current to decrease to its steady value of 2.29mA, as shown in graph (c). Consequently the anode potential also returns to its steady value of 243 volts, graph (d).

The third quarter of the cycle of voltage changes is represented in the region from 4 to 5 in each graph. During this time the signal voltage becomes 2.5 volts negative, the grid potential decreases from −10 to −12.5 volts, the anode current decreases from 2.29 to 1.35mA and the anode potential rises from 243 to 266 volts. During the fourth quarter of the cycle from 5 to 6 all voltages return to the steady state values at the beginning of the cycle. From the preceding discussion it may be seen that a variation of 5V in grid voltage produces a variation of 47V in the anode potential and thus voltage amplification has taken place.

The waveform of the anode current during the cycle of voltage changes is that of a *varying direct current*. It should be noted that the varying d-c is equivalent to a steady d-c with an a-c superimposed upon it. The *d-c component of anode current* is that current which flows through the valve when the grid signal passes through zero. It is the steady current which flows as a result of the steady grid bias voltage. The *a-c component of anode current* results from the a-c component of the grid voltage.

The waveform of anode voltage is likewise that of a *varying d-c voltage*. This waveform may also be thought of as the sum of a steady and an alternating component of anode voltage. The *d-c component of anode voltage* is the steady value of the anode voltage when no a-c signal is applied to the grid. The *a-c component of the anode voltage* results from the a-c component of the grid voltage. Since a rise in grid potential causes a fall in anode potential, the grid and anode voltage changes are 180° out of phase with each other.

2.2 Methods of obtaining bias. *Grid bias* is the initial steady potential between grid and cathode which determines the operating conditions of the valve.

In the type of amplifier described in the previous section the bias voltage is usually a little less than half the negative grid voltage necessary to cut off the flow of anode current. The signal then causes the grid to become less negative than the bias voltage during one half of the input cycle and more negative than the bias during the other half cycle. The peak of the signal voltage is always less than the bias voltage so that the grid potential does not become either positive to the cathode or sufficiently negative to cut off the flow of anode current.

Three methods of obtaining grid bias are commonly used in amplifier circuits.

A. Fixed Bias

Fixed bias is obtained from a separate source of voltage called the *C* supply as shown in Fig. 2-3. This voltage source may consist of a battery, a d-c generator or a rectified a-c power supply. It is connected with its positive terminal to

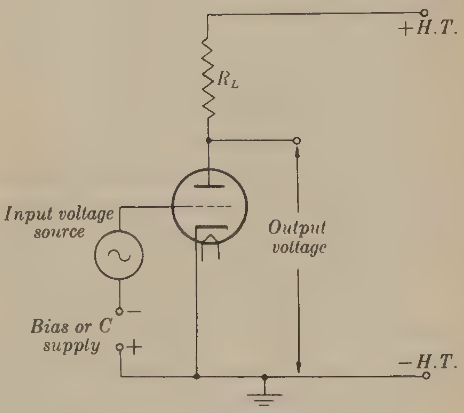


Fig. 2-3 Fixed bias using a battery.

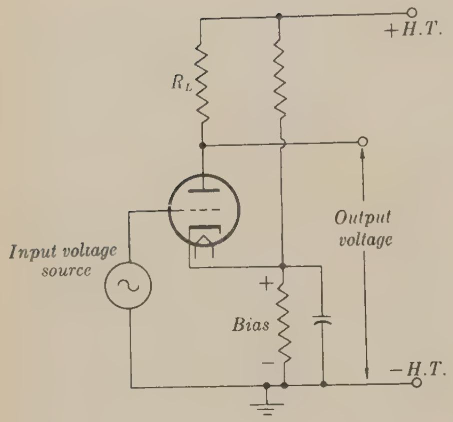


FIG.2-4 Fixed bias using a voltage divider.

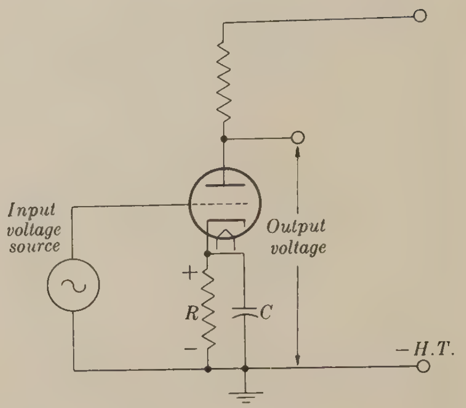


FIG.2-5 Cathode bias.

the cathode and its negative terminal to the grid. Fixed bias may also be obtained from a voltage divider across the H.T. supply (Fig. 2-4). Since the cathode connection is at a more positive potential on the voltage divider than the grid connection, the grid is negative with respect to the cathode.

B. Cathode Bias or Self-Bias

Cathode bias is obtained by connecting a *cathode bias resistor* between −H.T. and the cathode of the valve (Fig. 2-5). The steady anode current which flows through the valve when no signal is applied to the grid flows through this resistor. It develops a p.d. across the resistor with the polarity shown in the diagram. The cathode is connected to the positive end of the resistor while the grid is returned to the negative end through the source of the grid signal. The value of the bias is therefore equal to the p.d. across R, which is determined by the resistance of R and the current flowing through it.

In designing an amplifier circuit, the value of grid bias is determined from information given in the valve manual. The value of anode current which will flow when the bias and H.T. supply voltage are determined, is also given in the manual. The value of the cathode bias resistor which is required to produce the bias is then found by Ohm's Law.

Thus
$$R_c = \frac{E_c}{I_b}$$

where R_c = cathode bias resistance in ohms

E_c = bias voltage across R in volts

I_b = steady value of anode current in amperes.

If the valve is of the multi-grid type or if grid current flows, the value of I_b must be changed to the sum of the currents flowing from the cathode to all the other electrodes in the valve.

Example. What value of cathode resistance is required to provide a bias of 8 volts when the total cathode current is $4mA$?

$$R_c = \frac{8}{2\times 10^{-3}} = 4000$$

Therefore, the cathode resistance is 4000 ohms.

While the cathode bias resistor may be used by itself in the circuit, it is common to connect a *cathode by-pass capacitor* in parallel with the resistor. The purpose of the capacitor is best understood by considering the operation of the amplifier when an alternating voltage is applied to the grid.

During the positive half cycle of grid voltage, the current through the valve increases and so the p.d. across the cathode resistor increases. The bias therefore becomes greater which limits the current through the valve. That is, the anode current does not increase as much during the positive half cycle of grid voltage, as it would if the bias were maintained constant. Similarly, during the negative half cycle of grid voltage, the current through the valve decreases and so the bias is reduced. Therefore the anode current does not decrease as much as if the bias were maintained constant. As a result, the amplitude of anode current variations and hence of anode potential variations is less than that with constant bias. The result is that the amplification of the circuit is reduced. This effect is referred to as *degeneration* or *negative feedback*.

In order to maintain the bias constant and prevent degeneration, a capacitor is connected in parallel with the cathode resistor. The operation of this capacitor is as follows. When the grid is less negative with respect to the cathode on the positive half cycle of input voltage, the anode current which flows through the parallel *R*-*C* combination increases. Therefore the voltage across the *R*-*C* combination tends to increase. The capacitor is connected across the resistor to prevent the voltage across the resistor from increasing. That is, there is a small increase in the current through *R* and a relatively large current flow into the capacitor. Thus nearly all the increase in anode current is used in charging *C*. *C* is large enough so that it charges for the duration of the positive half cycle with very little change in the p.d. across it.

When the grid is made more negative to the cathode on the negative half cycle of input voltage, the anode current is reduced. This reduction in current tends to reduce the p.d. across the *R*-*C* combination. The reduction in p.d. is small because *C* discharges through *R*. Hence the reduction in anode current through *R* is nearly compensated for by the discharging current of *C* so that the actual drop in p.d. across *R* is small. *C* is large enough so that it can discharge through *R* for the duration of the negative half cycle with very little change in the p.d. across it. In terms of current components, *C* should have sufficient capacitance to *by-pass* the alternating component of anode current around *R* and permit only the d-c component to flow through *R*. In order to do this effectively, the reactance of the capacitor to the alternating current flowing through it should be less than one-tenth of the resistance of *R*. If currents at several frequencies are being amplified at one time, this reactance must be computed for the lowest frequency component which is to be by-passed.

Example. An amplifier is to amplify voltages at frequencies in the audio range from 200 c/s to 5000 c/s. The cathode resistor is 1000 ohms. Find the capacitance of the cathode by-pass capacitor which should be connected in the circuit.

Since the resistance is 1000 ohms, the reactance of the capacitor should be about 100 ohms at the lowest frequency of 200 c/s. Substituting in the formula

$$X_c = \frac{1}{2\pi fC}$$

$$100 = \frac{1}{6.28 \times 200C}$$

Therefore the required capacitance is approximately 8×10^{-6} farads or 8 microfarads. In practice, a 10-microfarad capacitor would be used.

C. Grid-Leak Bias

1. SHUNT TYPE.

Grid bias may also be obtained by placing a resistor and capacitor in the

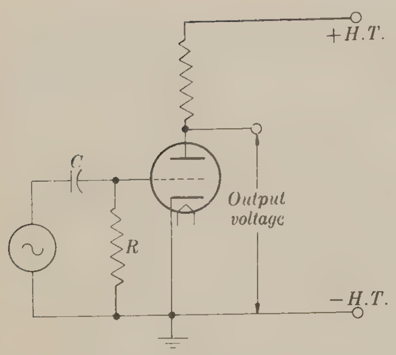


FIG. 2-6 Grid-leak bias, shunt type.

grid circuit as shown in Fig. 2-6. The grid capacitor is connected between the source of signal voltage and the grid. The resistor is connected between the grid and the cathode and is called the *grid-leak resistor.* Bias is established by developing a varying voltage across the grid-leak resistor, the d-c component of which makes the grid negative to the cathode.

In the absence of a signal voltage, there is no p.d. across R and hence there is no bias. During the first positive half cycle of input voltage, the grid is made positive to the cathode. The electrons which are attracted to the grid from the cathode constitute a flow of grid current. This current causes the capacitor to charge, the grid side being negative with respect to the signal side. During the negative half cycle of the signal voltage, the grid is made negative to the cathode so that grid current ceases to flow. The capacitor discharges a small amount through R. This develops a p.d. across R which makes the grid end negative with respect to the cathode end. The instantaneous grid voltage then equals the signal voltage plus the p.d. established across C by the flow of grid current as shown in Fig. 2-7. The values of C and R are so chosen that their time constant is long in comparison with the period of the signal voltage. Hence C loses only a small part of its charge during the negative half cycle of input voltage. During each succeeding cycle, the grid is driven positive to the cathode at the peak of the grid voltage swing when the signal voltage exceeds the voltage across C. C therefore charges further and the bias produced across R becomes larger. This condition is indicated during the first four cycles of the grid-signal voltage in Fig. 2-7. Thereafter the grid is driven positive for only a small time at the peak of each cycle. The grid current which flows during this time is just sufficient to re-establish the charge on C which is lost through R throughout the remainder of the cycle. The grid bias remains constant and its value is given by Ohm's Law,

$$E_{cc} = I_gR$$

where E_{cc} = grid bias in volts
I_g = d-c component of grid current flow in amperes
R = resistance of the grid-leak resistor in ohms.

In order to maintain a reasonably large bias with a small flow of grid current, the resistance of R is made very large and the capacitance of C very small. At the same time, the time constant of R-C must be large in comparison with the period of the signal voltage.

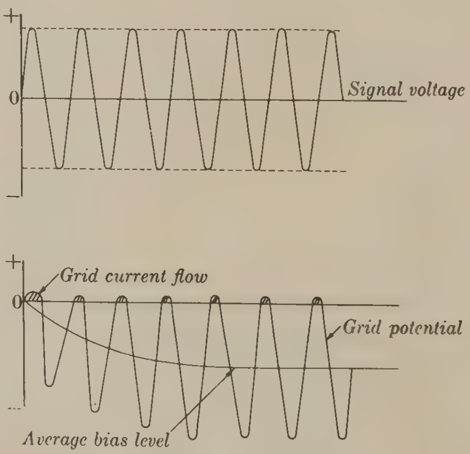


FIG. 2-7 Development of grid bias.

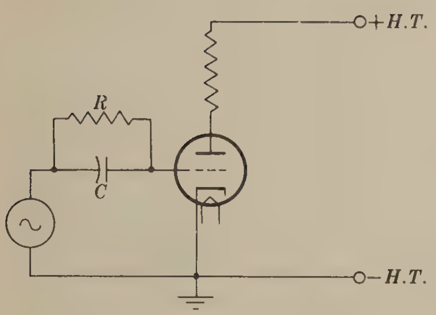


FIG. 2-8 Grid-leak bias, series type.

The amount of bias produced by the grid-leak arrangement is proportional to the amplitude of the grid signal. Since the grid-leak resistor is in parallel with the grid-cathode circuit of the valve, the bias is referred to as the *shunt type* of grid-leak bias.

2. SERIES TYPE.

A slightly different arrangement of circuit components (Fig. 2-8) results in the *series type* of grid-leak bias. The operation of the circuit during the establishment and maintenance of bias is similar to that of the shunt type. During a brief period at the positive peak of each input cycle, the grid is driven positive to the cathode. The capacitor charges, making the grid side negative with respect to the signal side. During the remainder of the cycle, *C* discharges somewhat through *R* maintaining the grid bias. As in the case of shunt bias, the time constant of *R-C* must be large so that the bias does not fall appreciably from one cycle to the next. This type of circuit arrangement also produces a bias which is proportional to the amplitude of the grid signal.

2.3 Static characteristics and dynamic characteristics. Two types of static characteristic curves of valves are described in Chapter I. They are the static e_b-i_b curves shown in Fig. 1-26 and the static e_c-i_b curves shown in Fig. 1-27. These characteristics are supplied by the manufacturer to describe the performance of the valve. When a valve is used in an amplifier circuit, an impedance called the *load* is connected between the anode and the positive terminal of the supply voltage. During normal operation of the amplifier, the grid voltage, anode current and anode voltage all vary at the same time. Consequently, the conditions in the valve, although represented by either family of static characteristics as a whole, are not represented by any one curve of the family. Moreover, the relationship between grid voltage and anode current, for a given valve, varies with different values of the load impedance. To describe the operation of the valve in an amplifier circuit therefore, it is necessary to draw a set of characteristic curves for the specific load impedance in use. Such a graph is the *dynamic characteristic* of the valve. There is just one *dynamic* e_c-i_b *characteristic* and one *dynamic* e_b-i_b *characteristic* for a given amplifier circuit.

In order to present these curves in their proper perspective, it is necessary to recall the method of obtaining the static characteristics referred to in Chapter I.

A. *The* e_c-i_b *Curves*

The static e_c-i_b characteristics are obtained by connecting the valve in the circuit of Fig. 2-9, with no load impedance in the anode circuit. With the anode voltage fixed, the grid voltage is varied in steps and the corresponding values of anode current are recorded. These values are plotted to give one curve of the family of e_c-i_b characteristics shown in Fig. 2-10.

The anode voltage is then set at another fixed value and a new set of values of grid voltage and anode current is obtained. These values are plotted on the same graph to give another curve of the complete family. A sufficient number of

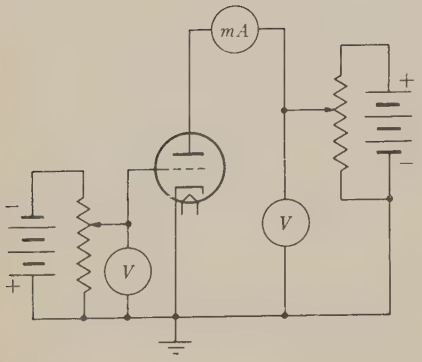


Fig.2-9 Circuit used to obtain static characteristics.

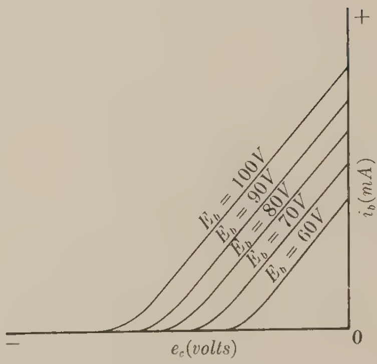


Fig.2-10 Static $e_c - i_b$ characteristics of a triode.

these curves is plotted to give a complete picture of the operation of the valve under normal operating voltages.

The dynamic characteristic is obtained in a similar manner but with a load impedance connected in the anode circuit (Fig. 2-11). The supply voltage is kept constant and the anode current is measured for various values of grid voltage. The dynamic characteristic shown in Fig. 2-12, is then obtained by plotting the values of grid voltage and anode current. The static characteristics are shown on the same graph for comparison with the dynamic characteristic. The latter is seen to lie across the static curves and not along or parallel to any one of them.

B. The $e_b - i_b$ Curves

The static $e_b - i_b$ characteristics of the valve are obtained by using the same circuit (Fig. 2-9) as that used to obtain the $e_c - i_b$ curves. Each curve shown in Fig. 2-13 is obtained by recording and plotting the values of anode voltage and anode current for a fixed grid voltage. Finally, the dynamic $e_b - i_b$ characteristic is obtained from the circuit of Fig. 2-14. The values of anode voltage and anode current are measured for various values of grid voltage. These are plotted to give the graph of Fig. 2-15 which is observed to be a straight line. This dynamic $e_b - i_b$ characteristic is called the *load line* of the amplifier. The static and dynamic $e_b - i_b$ characteristics are plotted on the

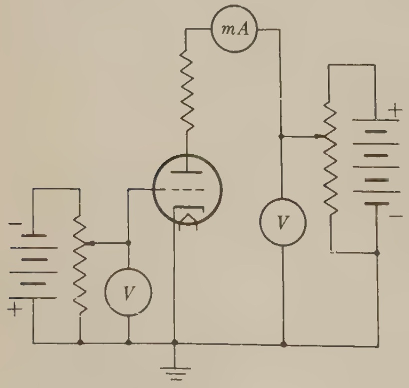


Fig.2-11 Circuit used to obtain dynamic $e_c - i_b$ characteristics.

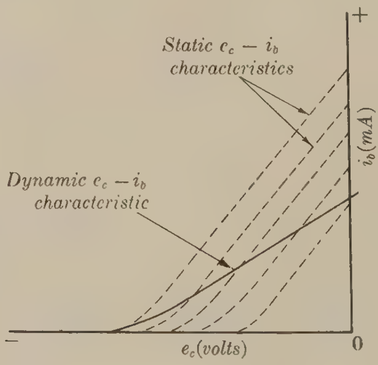


Fig.2-12 Static and dynamic $e_c - i_b$ characteristics.

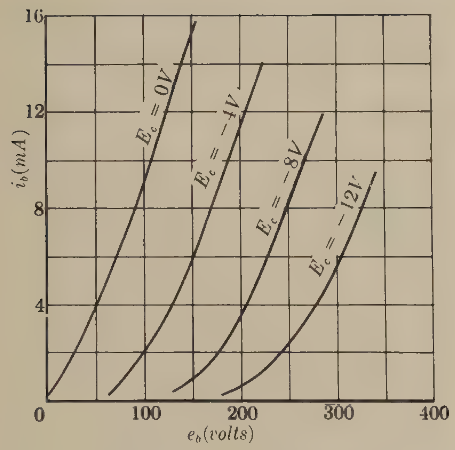


FIG.2-13 Static e_b-i_b characteristics.

same graph in Fig. 2-16. The load line is seen to lie across the static e_b-i_b curves.

C. Derivation of Dynamic Characteristics from Static Characteristics

The load line corresponding to any value of anode load resistance may readily be determined when the static e_b-i_b curves are available. Since the load line is known to be a straight line, only two points on it must be determined to fix its position. These are the zero anode current and zero anode voltage points and are obtained from the following considerations.

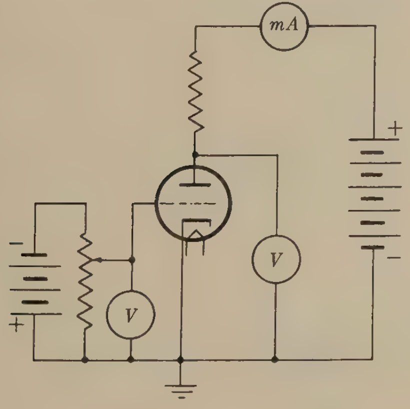


FIG.2-14 Circuit used to obtain dynamic e_b-i_b characteristics.

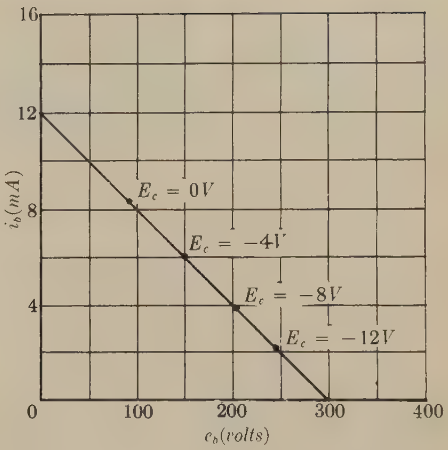


FIG.2-15 Dynamic e_b-i_b characteristic or load line.

If no current flows through the valve, a condition which exists when the grid potential is beyond the cut-off potential, there is no p.d. across the anode load resistor. The anode voltage is then at the value of the supply voltage. One point on the load line is therefore that at which the anode current is zero and the anode voltage is equal to the H.T. value. As the H.T. voltage is known, this point on the zero current axis is readily determined.

On the other hand, the maximum current that could flow through the valve

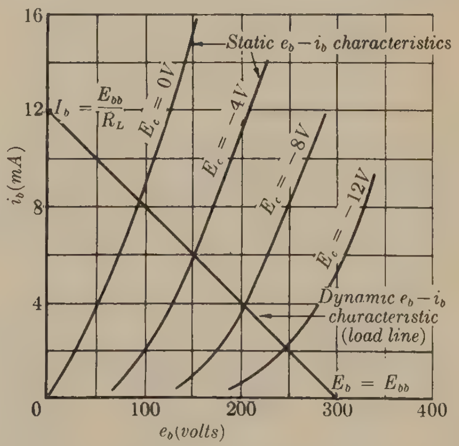


FIG.2-16 Static and dynamic e_b-i_b characteristics.

would be that which would develop the entire H.T. supply voltage across the anode load resistor. The anode voltage would then be zero. This value of anode current may be found from Ohm's Law as follows

$$I_b = \frac{E_{bb}}{R_L}$$

where I_b = maximum possible anode current in amperes
E_{bb} = supply voltage in volts
R_L = load resistance in ohms.

A plot of this value of anode current on the axis of zero anode voltage provides the second point on the load line. The straight line through these two points is then the load line.

The dynamic $e_c - i_b$ curve of an amplifier may be derived from the combined load line and static $e_b - i_b$ characteristics. A series of values of anode current and grid voltage are read from the points of intersection of the load line and the static $e_b - i_b$ curves of Fig. 2-16. These derived values are plotted to give the dynamic $e_c - i_b$ curve.

D. Application of the Characteristics in Circuit Analysis

The characteristic curves are used to determine the output voltage waveform

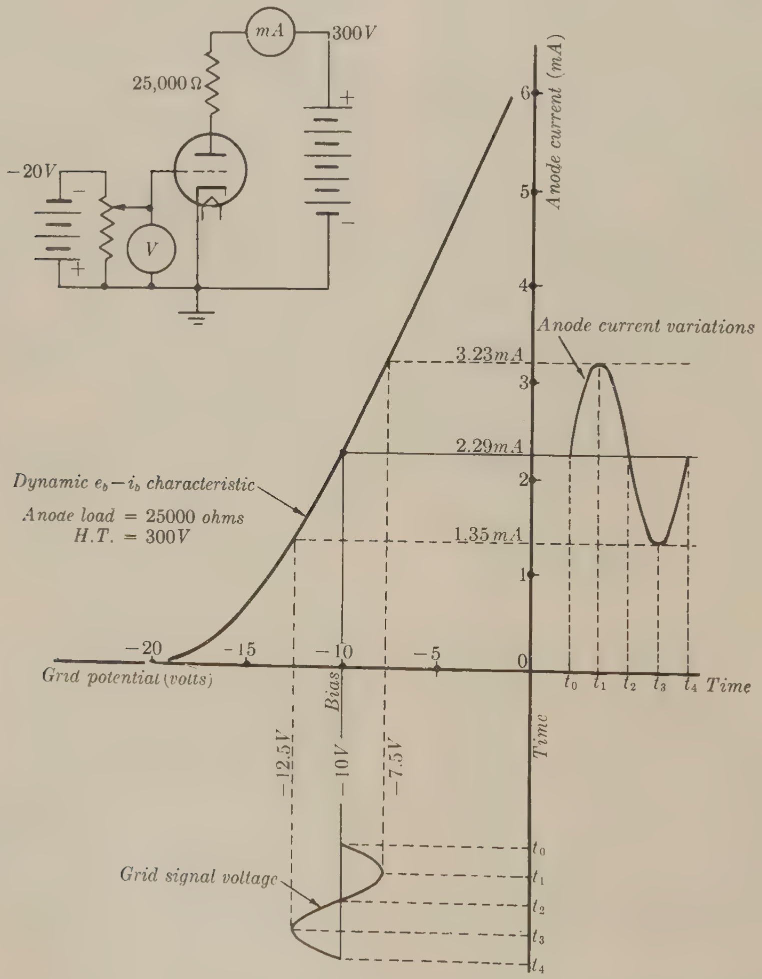


FIG. 2-17 Graphical representation of $e_c - i_b$ variations.

of an amplifier with a given input voltage waveform. There are three ways of doing this.

METHOD 1

The first method of determining the output voltage waveform makes use of the dynamic $e_c - i_b$ curve as shown in Fig. 2-17. The dynamic $e_c - i_b$ curve of the amplifier is obtained by using the circuit shown in the same diagram. It should be recalled that the dynamic characteristic refers to a specific value of anode load resistance and anode supply voltage as well as to a particular valve.

The grid bias voltage is represented by the point marked -10 volts on the grid potential axis to the left of the origin. A vertical line drawn through this point locates the operating point P on the $e_c - i_b$ curve. This line is also used as a *time axis* for the grid signal. One cycle of the sinusoidal grid-signal voltage is drawn on this axis, beginning at t_0 and ending at t_4. The displacement of any point on this curve from the vertical axis through the origin represents the instantaneous value of grid potential at the corresponding instant of time. The value of anode current corresponding to any value of grid potential is found by drawing a vertical line through the grid potential point to meet the $e_c - i_b$ curve. A horizontal line is then drawn from this point to the anode current axis and the corresponding value of anode current is determined. In particular, the value of anode current corresponding to a grid potential of -10 volts is seen to be $2.29 mA$. The horizontal line through the operating point is used as a time axis for plotting the variations in anode current produced by the cycle of grid voltage variations. To each point on the grid voltage curve, there corresponds a specific value of anode current. The complete set of anode current values plotted on the time axis gives rise to the complete cycle of anode current.

Having obtained the graphical picture of the operation of the circuit as shown in Fig. 2-17, the variations in anode voltage may be determined and the amplification calculated. The p.d. across the load resistor is found from Ohm's Law. This p.d. is subtracted from the supply voltage to give the value of anode voltage. For example, the anode current of the valve corresponding to -10 volts bias and no signal is $2.29 mA$. The p.d. across the 25,000 ohm load resistor is $\left(\frac{2.29}{1000} \times 25{,}000\right)$ or 57 volts. Hence the anode potential at the operating point equals $(300-57)$ or 243 volts. The table at the bottom of this page shows the values of grid potential, anode current, p.d. across the load resistor and the anode voltage respectively at times t_0, t_1, t_2, t_3 and t_4 of the grid voltage cycle.

As the grid potential varies from -7.5 volts to -12.5 volts, the anode potential varies from 219 volts to 266 volts. That is, a variation of 5 volts in grid potential produces a variation of 47 volts in anode potential. Hence the *amplification* of the circuit is $\frac{47}{5}$ or 9.4.

TIME	GRID POTENTIAL	ANODE CURRENT	P.D. ACROSS LOAD	ANODE POTENTIAL
t_0	-10 volts	2.29 mA	57 volts	243 volts
t_1	-7.5	3.23	81	220
t_2	-10	2.29	57	243
t_3	-12.5	1.35	34	266
t_4	-10	2.29	57	243

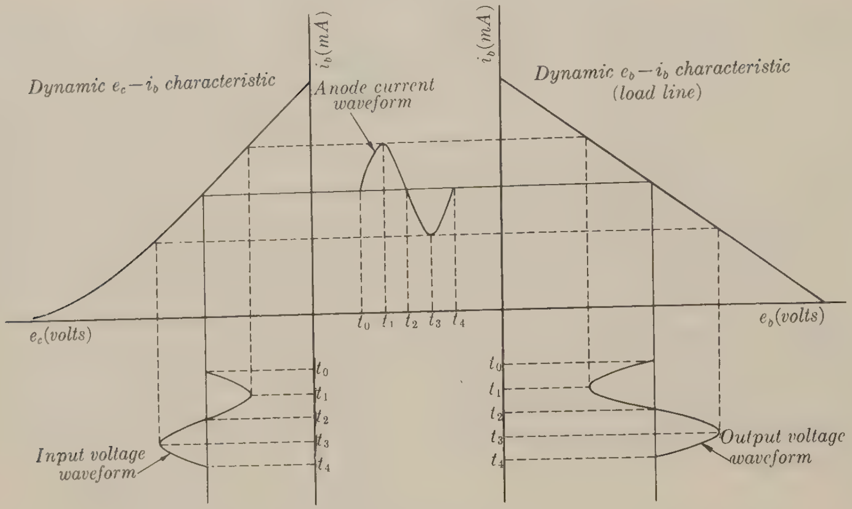


FIG.2-18 Graphical representation of $e_c-i_b-e_b$ variations.

METHOD 2

The second method of obtaining the output voltage waveform makes use of both the dynamic e_c-i_b and e_b-i_b curves as indicated in Fig. 2-18. This method is the same as the first method as far as finding the anode current waveform is concerned. The anode current waveform is then applied to the load line to find the output voltage waveform. The method of carrying out these operations is indicated by the dotted lines in Fig. 2-18. The dotted lines are projected upward from points of maximum and minimum grid voltage to the dynamic e_c-i_b curve and then horizontally to the load line. The heights of the horizontal dotted lines represent the values of the anode current corresponding to the maximum and minimum values of grid voltage. The times corresponding to these values of grid voltage and hence to those of anode current may be read from the input voltage curve. Thus the anode current waveform is plotted on a time base between the dynamic mutual characteristic and the load line. The dotted lines are now projected down from the load line. The displacements of the lines to the right of the i_b axis of the load line represent the values of anode voltage at the times corresponding to the dotted lines. An output voltage waveform therefore, is plotted on a time base below the load line graph as indicated. It should be noted that the anode current waveform need not be drawn to find the output voltage.

METHOD 3

A third method of obtaining the output voltage waveform using the load line is illustrated in Fig. 2-19. Because each static e_b-i_b curve corresponds to a different grid voltage, the load line may be used as a grid voltage axis for the graph. The input voltage waveform is applied to the graph along a time axis perpendicular to the load line. The output voltage is found by projecting dotted lines from the points of maximum and minimum grid voltage upon the load line. Dotted lines are then projected vertically downward to the anode voltage axis indicating the values of minimum and maximum anode voltage. The anode current waveform is plotted by projecting dotted lines on the anode current axis from the points on the load line which represent the maximum and minimum grid voltage. This method has the

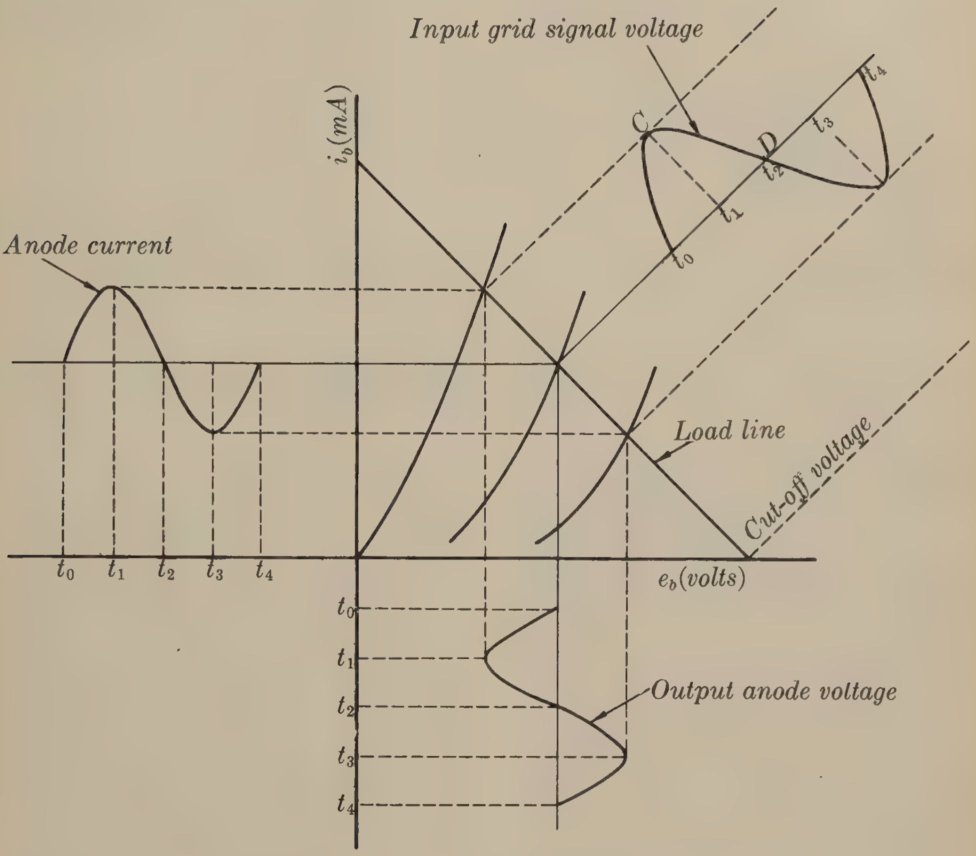


FIG.2-19 $e_c - i_b - e_b$ variations plotted on load line.

advantage of requiring fewer graphs than the previous method.

2.4 Determination of amplification from the equivalent circuit. The graphical methods described in the preceding section are of the utmost value in understanding the operation of an amplifier and in predicting its performance. In particular, the amplification or *gain*, *A*, of an amplifier may be determined by means of these graphs. The amplification is the ratio of the output voltage to the input voltage. It may also be expressed in terms of the valve factors and the circuit constants by a consideration of the equivalent circuit of the amplifier. The actual circuit and the equivalent circuit are shown in Fig. 2-20 and Fig. 2-21 respectively. In order to appreciate their equivalence, the meaning of amplification factor is recalled. The amplification factor μ is equal to a very small change in anode voltage $\triangle e_b$, divided by the change in grid voltage $\triangle e_c$ which would produce the same change in anode current $\triangle i_b$ as was produced by $\triangle e_b$. Expressed as a formula, this becomes

$$\mu = \frac{\triangle e_b}{\triangle e_c} \text{ or } \triangle e_b = \mu \triangle e_c$$

So far as the circuit is concerned, the valve is the equivalent of an a-c generator which produces a voltage of magnitude $\mu \triangle e_c$ in series with the a-c resistance of the valve r_a, and the load resistance R_L. The generator causes the current $\triangle i_b$ to flow through the resistances r_a and R_L.

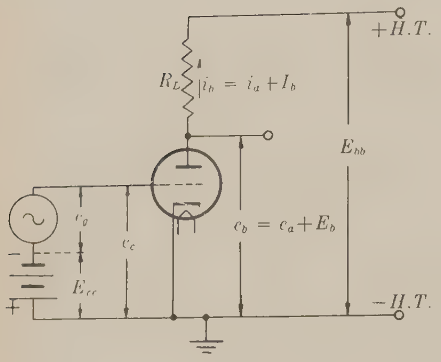


FIG.2-20 Amplifier circuit showing voltage and current symbols.

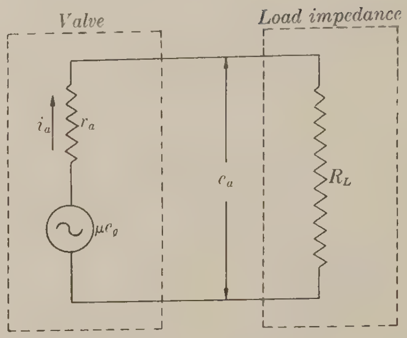


FIG.2-21 Equivalent circuit of a resistance-loaded amplifier.

This concept is illustrated by the circuit of Fig. 2-21 in which the valve and the external circuit are separately enclosed in dotted lines. An analysis of the operation of the equivalent circuit is somewhat simpler than that of the original circuit because the d-c components of current and voltage in both the input and output circuits are eliminated. These components are of no consequence in determining the gain of the circuit as an a-c voltage amplifier.

The change in anode current $\triangle i_b$ which is caused by the change in grid voltage $\triangle e_c$ is given by Ohm's Law as follows:

$$\triangle i_b = \frac{\mu \triangle e_c}{R_L+r_a}$$

The change in voltage produced across R_L caused by this changing current is also given by Ohm's Law

$$\triangle e_b = \triangle i_b \times R_L$$

Substituting for $\triangle i_b$ gives

$$\triangle e_b = \frac{\mu \triangle e_c}{R_L+r_a} \times R_L$$

In an amplifier the changes in current and voltage are the alternating components caused by the alternating input voltage. Therefore $\triangle e_c$, $\triangle i_b$ and $\triangle e_b$ may be replaced in the expressions by e_g, i_a and e_a respectively.

Hence the gain $A = \frac{e_a}{e_g}$

$$= \frac{\dfrac{\mu e_g \times R_L}{R_L+R_a}}{e_g}$$

$$\therefore A = \frac{\mu R_L}{R_L+r_a}$$

where A = stage gain
μ = amplification factor of the valve
R_L = effective resistance of load in ohms
r_a = anode resistance of the valve in ohms.

The gain of the amplifier shown in Fig. 2-20 is calculated as follows:

$$\mu = 20$$
$$r_a = 7700\Omega$$
$$R_L = 8000\Omega$$
$$\therefore A = \frac{20 \times 8000}{7700+8000}$$
$$= \frac{160{,}000}{15{,}700}$$
$$= 10.2$$

Therefore the amplification is 10.2.

The values of μ and r_a were obtained from the valve manual. The result given by this equation agrees closely with that found by the graphical method from which the value of the gain is 9.4. Thus the need of drawing the graph to determine the gain is eliminated.

If R_L is increased to 15,000 ohms, the gain becomes

$$A = \frac{20\times 15{,}000}{15{,}000+7700} = \frac{300{,}000}{22{,}700} = 13.2$$

Thus, for a resistive load, the gain may be increased by:

(i) selecting a valve with a high μ

(ii) making R_L very large compared with r_a.

If R_L is made very large, then the value of $\frac{R_L}{R_L+r_a}$ becomes nearly equal to unity. In this case, the gain approaches a value equal to the amplification factor of the valve. In practice it is possible to obtain a gain almost equal to μ. However, when R_L is increased, the d-c voltage drop, I_bR_L, across it is increased and the anode voltage E_b is reduced. This leads to inefficient operation of the valve. Hence, a larger H.T. supply is necessary to provide the required anode voltage. Usually, the value of R_L which is selected for the amplifier is a compromise reached after considering several conflicting factors such as the H.T. supply voltage available, the gain required, the amplification factor and the a-c resistance of the valve.

2.5 Two-stage a-f voltage amplifiers. An amplifier is frequently required to provide a higher gain than that possible from a single stage of amplification. The required degree of amplification may be obtained by connecting two or more amplifier stages in series or *cascade.* The output voltage from the first stage is coupled to the grid of the second stage. The circuit used to provide the load impedance of the first amplifier and pass on the output voltage to the second amplifier is called the *inter-valve* or *inter-stage coupling circuit.* Three types of coupling circuits are used with a-f voltage amplifiers. They are:

(i) resistance-capacitance coupling

(ii) inductance-capacitance coupling

(iii) transformer coupling.

A complete understanding of the coupling circuit is necessary in order to appreciate the operation of the amplifier as a whole.

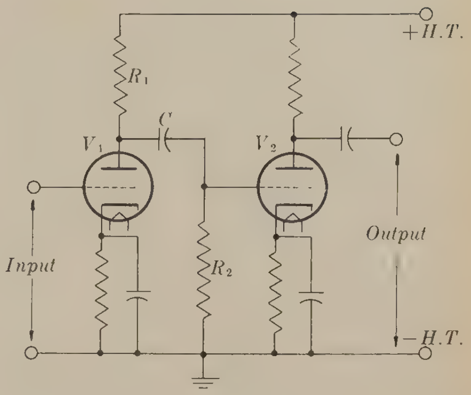


FIG.2-22 Resistance-capacitance coupled amplifier.

A. *Resistance-Capacitance Coupling*

A two-stage voltage amplifier with resistance-capacitance coupling between stages is shown in Fig. 2-22. Triode valves V_1 and V_2 are used in identical circuits. Bias is provided by a resistance-capacitance network in the cathode circuit of each valve. The a-c voltage to be amplified is applied between grid and cathode of the first valve. The amplified output voltage of this stage is developed across the anode load resistor R_1. C is the *coupling capacitor* which couples the a-c output voltage of the first stage to the grid of the second stage. In addition this capacitor blocks the d-c voltage at the anode of V_1 from the grid of V_2. R_2 is the *grid-leak resistor* of the second stage.

The operation of the coupling circuit is best understood by referring to the voltage waveforms of Fig. 2-23. The graphs show two phases in the operation of the circuit. During the time interval indicated by XY, no a-c signal is applied to the grid of V_1, while in the interval

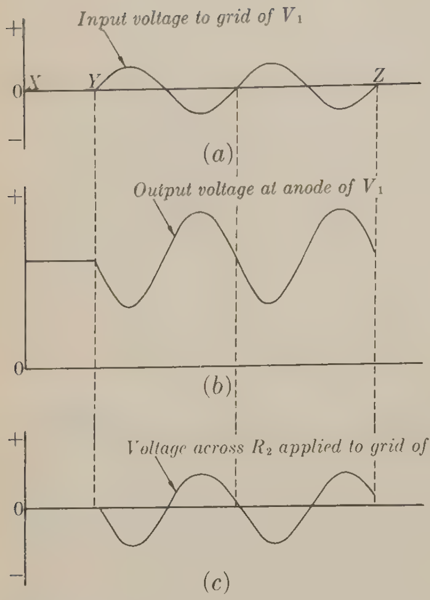


FIG. 2-23 Voltage waveforms produced by R-C coupling.

YZ an a-c voltage of sinusoidal waveform and constant amplitude is applied as the input to V_1.

When no signal is applied to the grid, the current through the valve is a steady d-c. This current develops a d-c voltage across R_1 causing the anode potential of V_1 to remain at a steady value which is somewhat less than the H.T. supply voltage. The coupling capacitor C charges through R_1 and R_2 until the p.d. across it equals the anode potential. Thereafter, the voltage across C remains constant and no p.d. exists across R_2. Hence the grid potential of V_2 is maintained at zero. It is important that there be very little leakage of electrons through C as this would cause a positive potential on the grid of V_2 in opposition to the cathode bias.

During the interval from Y to Z, an a-c signal is applied to the grid of V_1. The input voltage shown in (a) causes fluctuations in current through the valve which in turn causes the anode potential to vary as shown in (b). The variations in anode voltage are then applied across C and R_2 in series. C blocks the d-c potential and most of the a-c component of the anode voltage appears across R_2 as shown by the waveform in (c).

The values of C and R_2 are chosen after a consideration of two factors. Since C and R_2 in series are in parallel with V_1, the total impedance of C and R_2 should be large in order to prevent a short-circuiting effect on V_1. This factor suggests that the value of R should be very large and that the capacitance of C be small. However the minimum size of C is determined by the frequency of the a-c signal. C is chosen to be sufficiently large so that most of the a-c voltage at the anode of V_1 is developed across R_2 and only a small part of it is lost across C. To accomplish this the resistance of R_2 is at least 10 times as great as the reactance of C at the frequency of the signal. Under these conditions also, the a-c grid voltage of V_2 is almost in phase with the anode voltage of V_1 and 180° out of phase with the grid voltage of V_1. In practice the coupling capacitor in resistance-capacitance coupled a-f amplifiers is of the order of $.001\mu F$ to $.1\mu F$ and the grid-leak resistor 100,000 ohms to $1 M\Omega$.

The a-c voltage applied to the grid of V_2 is normally smaller in amplitude than the grid bias applied to V_2. Hence, the grid is negative to the cathode during all parts of the cycle and no electrons are attracted to the grid. The coupling capacitor therefore charges and discharges only through the grid-leak resistor R_2. If the grid-leak resistor is open-circuited, operation of the amplifier may appear normal for a time. Eventually however, electrons striking the grid, in spite of its negative bias, accumulate on C causing the bias to increase. If this continues, the coupling capacitance C may charge to a higher d-c voltage than the steady component of anode voltage of V_1. Hence the bias on the second valve

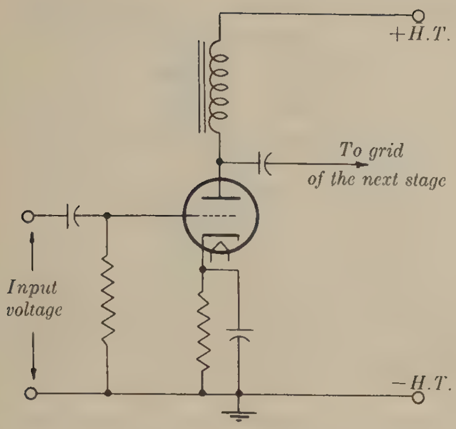


FIG.2-24 Inductance-capacitance coupling.

increases until it cuts the valve off. This is called *grid blocking*. It may even occur if R_2 is present in the circuit but has an excessively large resistance. In this case blocking is only temporary until the electrons which have accumulated on C have had time to leak off and the bias is sufficiently reduced to permit current to flow again in V_2. Intermittent blocking of this type is called *squegging*.

B. Inductance-Capacitance Coupling

It is possible to replace the anode load resistor of an a-f voltage amplifier by an iron-core inductor or the primary of an a-f transformer. Such a coupling circuit is referred to by the term *impedance coupling*. An a-f voltage amplifier which employs *inductance-capacitance* or *choke-capacitance coupling* is shown in Fig. 2-24. The operation of the grid and cathode circuits is the same as that described for resistance-capacitance coupled amplifiers. The voltage to be amplified is applied between grid and cathode of the valve. The variations in cathode-to-anode electron flow caused by the a-c grid signal, develop a varying voltage across the inductance in the anode circuit. The a-c voltage produced at the anode is then applied to the following stage across the coupling capacitor and grid-leak resistor in series.

This type of load has the advantage that the d-c resistance of the inductor is usually very small, of the order of a few hundred ohms. Consequently, the d-c voltage developed across the load is also small and the steady anode potential is close to the potential of the H.T. supply. A lower value of supply voltage may therefore be used than in the case of a resistance load to produce the same anode potential. The a-c impedance of the load may be made very large by making the inductance large. The reactance of the load is given by the formula

$$X_L = 2\pi f L$$

where X_L = reactance of the load in ohms
f = frequency of the a-f voltage which is being amplified in c/s
L = inductance of the load in henries.

The equivalent circuit of an amplifier with impedance loading is shown in Fig. 2-25. It is similar to that of the resistance-loaded amplifier of Fig. 2-21, except that the load contains two components R_L and L_L. R_L is the resistive component of the inductor while L_L is the inductive component.

When the load is inductive, the phase angle between the grid voltage and the anode voltage of the valve is no longer 180°. The phase angle depends upon the relative values of inductance and resistance in the inductor. It also depends

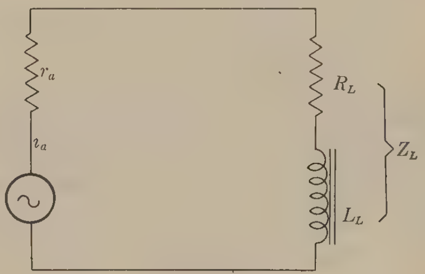


FIG.2-25 Equivalent circuit of impedance-loaded amplifier.

upon the relative values of the load impedance and the a-c resistance of the valve. The departure from 180° is greatest when the inductive reactance of the load is very large in comparison with its resistance and when the impedance of the load is small in comparison with the a-c resistance of the valve. Under these conditions the phase angle between the input and output voltages approaches 270°. The formula for calculating the gain of an impedance-loaded amplifier is somewhat different from that of the resistance-loaded amplifier which was derived in section 2.4 as

$$A = \frac{\mu R_L}{r_a + R_L}$$

In the case of an impedance load the total resistance in the equivalent circuit is $R_L + r_a$ which is in series with the inductive reactance X_L. Since the total resistance and the reactance are vectorially at right angles the magnitude of the total impedance of the circuit is given by

$$\sqrt{(r_a + R_L)^2 + X_L^2}$$

Further, the magnitude of the impedance of the inductor is $\sqrt{R_L^2 + X_L^2}$.

Thus the gain is given by the formula

$$A = \frac{\mu \sqrt{R_L^2 + X_L^2}}{\sqrt{(r_a + R_L)^2 + X_L^2}}$$

where A = stage gain
μ = amplification factor of valve
R_L = resistance of inductor in ohms
X_L = inductive reactance of inductor in ohms
r_a = anode resistance of valve in ohms.

If the resistance of the inductor is negligible compared to its reactance and to the anode resistance of the valve, which is usually the case in practice, this formula becomes

$$A = \frac{\mu X_L}{\sqrt{r_a^2 + X_L^2}}$$

where A = stage gain
μ = amplification factor of valve
and X_L = inductive reactance of inductor in ohms
r_a = anode resistance of valve in ohms.

This formula may be written in the form

$$A = \frac{\mu}{\sqrt{1 + \left(\frac{r_a}{X_L}\right)^2}}$$

It should be noted that the gain of the amplifier depends upon the ratio of r_a to X_L. If the load reactance is very large in comparison with the a-c resistance of the valve, then $\frac{r_a}{X_L}$ has a small value. The gain of the amplifier is then almost as large as the amplification factor μ of the valve. On the other hand, if the load reactance is small in comparison with the a-c resistance of the valve, the gain is considerably smaller than the amplification factor. Further, since the reactance of the load depends upon the frequency of the a-c voltage being amplified, the gain varies at different frequencies, becoming greater as the frequency increases.

To sum up, the gain of an impedance-loaded amplifier is increased

(i) by making the inductive reactance of the load larger in comparison with its resistance
(ii) by making the reactance of the load larger in comparison with the a-c resistance of the valve
(iii) by increasing the frequency of the signal voltage within certain limits.

The principle of resonance may also be used to advantage in the inductance-loaded amplifier. If the inductance of the load is made to resonate with the self-capacitance of the inductor and the interelectrode capacitance of the valve, the load may be made a parallel resonant circuit at the frequency of the signal which is being amplified. Under these conditions the load impedance is a maximum and is approximately resistive.

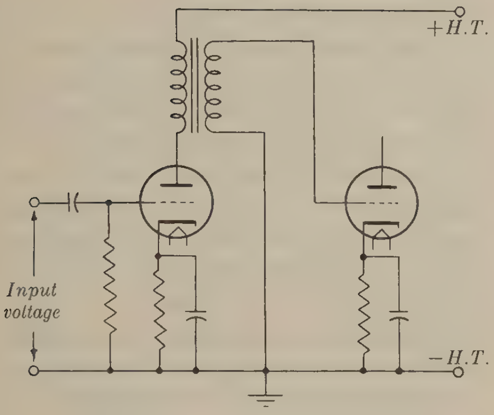


FIG.2-26 Transformer-coupled amplifier.

Consequently the gain of the amplifier is very high.

In most a-f voltage amplifiers, inductance-capacitance coupling has no advantages over transformer coupling which is discussed below. Furthermore, inductance-capacitance coupling is more bulky than resistance-capacitance coupling. For these reasons, it is not now often used. The development of high-quality transformers and high-mu valves has rendered inductance-capacitance coupling obsolete in a-f voltage amplifiers.

C. Transformer Coupling

An a-f voltage amplifier which employs transformer coupling between stages is shown in Fig. 2-26. The primary winding of the coupling transformer is connected in the anode circuit of the first valve. The secondary circuit of the transformer is connected between grid and cathode of the following stage.

The a-f voltage applied to the grid of the valve causes variations in the anode current which flows through the primary of the coupling transformer. The a-c component of the anode current sets up a varying magnetic field about the primary circuit which links the turns of the secondary and induces in it a corresponding a-c voltage.

As in the case of the resistance-loaded and inductance-loaded amplifiers, the effective impedance of the transformer primary must be large in comparison with the anode resistance of the valve. In this way the gain of the stage may be made to approach the amplification factor of the valve. In order to achieve this condition, comparatively low-mu triode valves are used. Typical values of μ range from 8 to 20 and values of a-c resistance from 7500 to 15,000 ohms are common.

A further gain is provided by a step-up transformer ratio of about 1:3. Such a ratio usually provides a reasonable match between the required a-c load impedance in the anode circuit and the higher impedance of the grid circuit to which the secondary is connected.

As in the case of inductance loading, the transformer has the advantage of presenting a very small d-c resistance in the anode circuit of the valve. Consequently the d-c voltage drop in the primary is small and a lower value of supply voltage may be used than with a resistance-loaded amplifier.

2.6 Requirements of audio-frequency amplifiers. In order to appreciate the requirements which must be met by the audio-frequency amplifier in the reproduction of audio signals, it is necessary to understand the response of the ear to sounds of varying pitch, loudness and quality.

A. Effect of Pitch

The pitch of a sound is the auditory effect of frequency. When the note of lowest pitch on a piano is played by striking the key at the extreme left end of the keyboard, the piano string is made to vibrate about 27 times per second. The string sets up vibrations in the sounding board and in the surrounding air, which are carried to the ear of the listener. The sound waves entering the ear set the ear drum in vibration at the frequency of the string. These vibrations are transmitted to the nerve endings in the inner ear and are interpreted by the brain as a sound of low pitch. If now the key at the extreme right end of the piano keyboard is struck, a string is set in vibration at a frequency of 4096 vibrations per second. The diaphragm of the ear is made to vibrate at the same frequency and the brain interprets the sound as one of high pitch. The range of human hearing is not limited to that of the piano, of course; most people are able to hear sounds over the frequency range from 16 c/s to about 16,000 c/s.

B. *Effect of Loudness*

The range of human hearing is limited not only as to pitch, but also as to intensity. The lowest sound intensity that can be heard by the ear is referred to as the *threshold of hearing* and is represented by the lower boundary of the chart in Fig. 2-27. The top boundary of this chart is called the *threshold of feeling* and sounds of greater intensity become painful to the ear. Tests indicate that the ear is most sensitive to sounds of about 2000 c/s and that its response is fairly uniform over the range 800 c/s to 5000 c/s. The sensitivity falls off progressively for frequencies below 800 c/s and above 5000 c/s as indicated in the chart of Fig. 2-27.

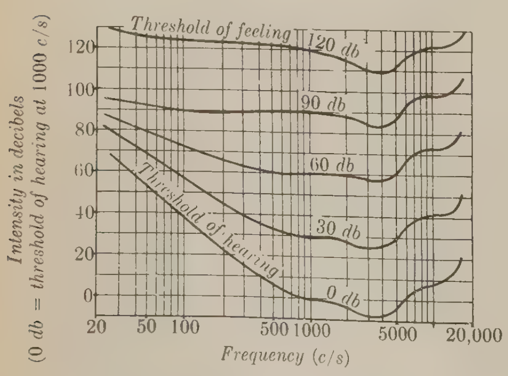


Fig. 2-27 Range of human hearing.

In addition to the non-uniformity in the response of the ear indicated above, the ear has a logarithmic response to intensity of sound. That is, the apparent loudness of sound is not proportional to the intensity of the source but to the logarithm of the intensity. In order to increase the loudness of a sound to double its former value, the intensity must be increased ten times.

The gain of an amplifier is measured in *decibels* above or below a fixed reference level which is usually the threshold of hearing. One decibel is equal to 1/10 of a *bel*. The number of bels equals the logarithm of the ratio of the two levels of intensity or power. The number of decibels is therefore equal to ten times the logarithm of the ratio of the two power levels.

That is, $\text{No. of bels} = \log \frac{P_1}{P_2}$

$$\text{No. of decibels} = 10 \log \frac{P_1}{P_2}$$

where P_1 = the larger power in watts

P_2 = the smaller power in watts.

In the case of an amplifier in which the output power is greater than the input power, the gain is expressed in positive decibels (+*db*). If the output power is less than the input power, the attenuation is expressed in negative decibels (−*db*).

Thus to compare two power levels of 3 watts and 30 watts in bels and decibels,

$$\text{No. of bels} = \log \tfrac{30}{3} = \log 10 = 1$$

$$\text{No. of decibels} = 10 \log \tfrac{30}{3} = 10.$$

That is, an increase in power of 10 times is equivalent to an increase of 1 bel or 10 decibels and would sound twice as loud to the ear. The boundaries in Fig. 2-27 marked 0, 30, 60, 90, 120 decibels therefore represent loudness levels in the ratios 0: 3: 6: 9: 12. The logarithmic relationship between the intensity of the stimulus and the physiological response is known as the *Weber-Fechner Law*.

C. Effect of Quality

It is convenient for purposes of analysis to consider the effect of sounds of individual frequencies when investigating the operation of the ear, an amplifier, microphone, loudspeaker, or any other audio device. It should be noted that nearly all sounds which reach the ear are of a very complex nature and consist of a blend of many sounds which differ in loudness and pitch arriving at the ear simultaneously. The particular combination of these components gives the sound its *quality*. It is the slight differences in quality of speech which makes it possible to distinguish one person's voice from another. Similarly, differences in quality of the sounds produced make it possible to distinguish between a piano and a violin. Both instruments may be producing the same fundamental tone but differ in quality because of the difference in the number and prominence of the harmonics mixed with the fundamental. In the same way, the quality of an amplifier is dependent upon its ability to amplify sounds of a wide range of frequencies and in their proper proportions. Moreover, it may be seen from Fig. 2-27 that if the level of sound volume produced by an amplifier is to be increased from say 30*db* to 60*db*, the amount of gain required in the middle frequency range is not the same as that in the low- and high-frequency ranges. Conversely, if an amplifier increases the volume of all frequencies by the same amount the quality of the reproduction may appear to change as far as the ear is concerned.

Speech frequencies usually fall within the limits 200 c/s to 3000 c/s. If either the low frequencies or the high frequencies are not produced, the signal becomes less intelligible. Tests show, however, that voice signals still retain 90% of their intelligibility when all frequencies below 900 c/s are eliminated. Similarly, all frequencies above 3000 c/s can be eliminated and 90% of the intelligibility is maintained. In each of these cases, however, the reproduction sounds unnatural. If the middle-frequency range is narrowed beyond these limits, intelligibility falls off rapidly.

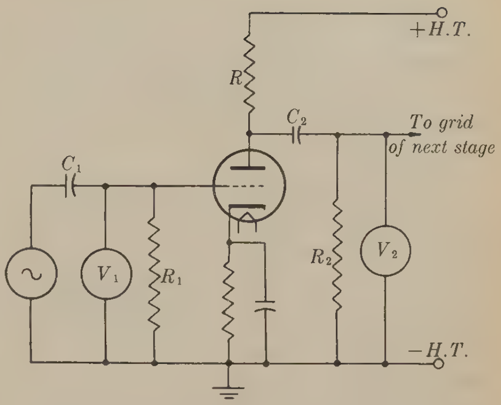


FIG. 2-28 Circuit used to obtain frequency response of an amplifier.

2.7 Frequency response of an amplifier. The term *frequency response* when applied to an amplifier refers to the ability of the circuit to amplify voltages over a given range of frequencies. If a circuit provides amplification over a comparatively small range of frequencies, it is said to have a *narrow* or *selective* frequency response. If it amplifies uniformly over a very wide range of frequencies it is said to have a *wide* or *flat* response. The significance of these terms is best understood by a consideration of the *frequency-response curve* of an amplifier.

An experimental circuit used to determine the frequency response is shown in Fig. 2-28. The amplifier with its input coupling circuit C_1R_1 and output coupling circuit C_2R_2 is provided with normal supply voltages. A sinusoidal input voltage of variable frequency is applied by a suitable signal generator connected in the grid-cathode circuit. An a-c voltmeter V_1 is connected across R_1

to measure the amplitude of the input voltage. A second a-c voltmeter V_2 is connected across R_2 to measure the output voltage. From the readings of these two voltmeters, the gain may be determined.

The amplitude of the input voltage is maintained constant and at a normal level for the particular amplifier throughout the experiment. The frequency of the input is then varied in steps from the lowest to the highest frequency which the amplifier is designed to handle. The reading of V_2 is noted at each of these frequency settings and a table of values of gain and frequency is recorded. From these values, a graph is made using frequency as abscissae and gain as ordinates. Such a curve is called the frequency-response curve of the amplifier. The graph of Fig. 2-29 illustrates the type of response curve obtained with a resistance-capacitance coupled a-f voltage amplifier. The central portion of the graph indicates the region over which the gain of the amplifier remains substantially uniform. The term *flat response* applied to an amplifier, therefore, refers to *uniform amplification*. The frequencies included in this range are called the *mid-frequencies*. If this range is comparatively large, the amplifier is said to have a *wide* frequency response. If the mid-frequency range is small the amplifier has a *narrow* or *selective* response. Frequencies below the mid-frequency range are called the *low frequencies* while those above the mid-frequency range are called the *high frequencies*.

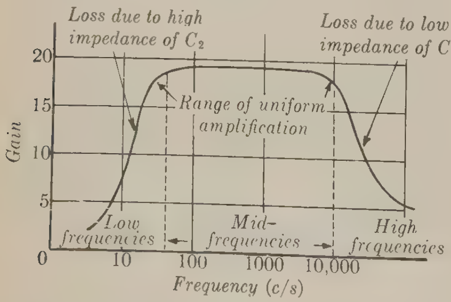


FIG. 2-29 Frequency-response curve of an audio-frequency amplifier.

The most desirable shape for the frequency-response curve depends upon the purpose for which the amplifier is designed. If the circuit is to be used for the amplification of radio-telephony signals, it should have a flat response over the range of voice frequencies. These frequencies generally fall within the limits 200 c/s to 3000 c/s.

For high fidelity reproduction of r-t signals and for music, the flat portion of the response curve should be extended to cover the range from 20 c/s to 20,000 c/s. This range includes all the frequencies produced by musical instruments from the lowest notes of the bass horn to the shrill sounds in the upper registers of the piccolo and violin. Moreover, this range of frequencies encompasses the entire range of human hearing.

The factors which affect the shape of the frequency-response curve of a resistance-capacitance coupled a-f voltage amplifier are best understood by reference to the equivalent circuit shown in Fig. 2-30. This circuit is similar to the equivalent circuit of Fig. 2-21 but includes in addition the coupling capacitor C_2 and the grid-leak resistor R_2 of the following stage. The voltage developed across this resistor is the input signal to the following stage. In addition the diagram shows the input and output capacitances

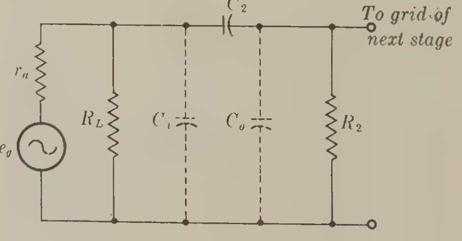


FIG. 2-30 Equivalent circuit of an a-f voltage amplifier.

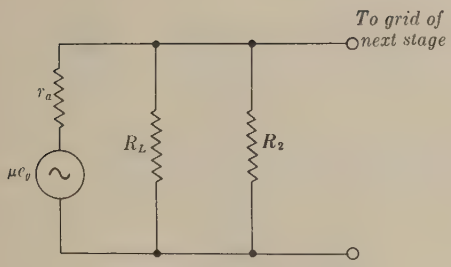


FIG.2-31 Equivalent circuit of an a-f amplifier for mid-frequencies.

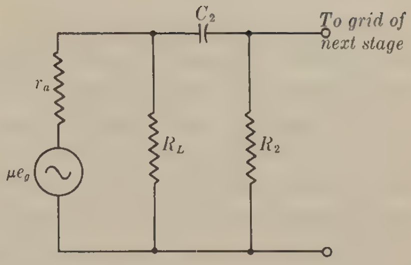


FIG.2-32 Equivalent circuit of an a-f amplifier for low frequencies.

of the valve C_i and C_o respectively. These capacitances affect the shape of the response curve in the high-frequency regions.

An amplifier is usually designed to give maximum gain at the mid-frequencies. It is therefore convenient to begin an analysis of the circuit operation at these frequencies. The coupling capacitor C_2 which is connected in series with R_2 across R_L is chosen so that its reactance at the mid-frequencies is small in comparison with the resistance of R_2. It is effectively an a-c short circuit between the anode of the first valve and the grid of the second valve and may be omitted from the equivalent circuit diagram. Also, at the middle and low frequencies, the input and output capacitances of the valve may be neglected. Since the reactance of a capacitor is inversely proportional to the frequency of the voltage applied to it, C_i and C_o present very large impedances in parallel with R_L and R_2 and have practically no shunting effect on the signal. The equivalent circuit of the amplifier therefore reduces to that shown in Fig. 2-31 for the mid-frequencies.

The decrease in gain at low frequencies is due to the increase in the reactance of the coupling capacitor as the frequency decreases. As a result, a larger fraction of the output voltage appears across the capacitor C_2 and a smaller fraction of the voltage appears across the grid-leak resistor R_2 of the next stage. Since only the latter voltage forms the input to the next stage, the over-all gain falls off rapidly in the low-frequency region. The equivalent low-frequency circuit is shown in Fig. 2-32. If it is desired to extend the flat portion of the response curve toward the low-frequency end, the value of C_2 must be increased.

At high frequencies, the input and output capacitances of the valve together with stray wiring capacitance reduces the gain. These capacitances are so small and their reactances so large at low frequencies that their effects may be neglected. The interelectrode capacitances which determine the input and output capacitances of a triode valve are illustrated in the diagram of Fig. 2-33. The input capacitance C_i is due to the capacitances C_{gk} between grid and cathode and C_{ga} between grid and anode. These two capacitances between the grid and

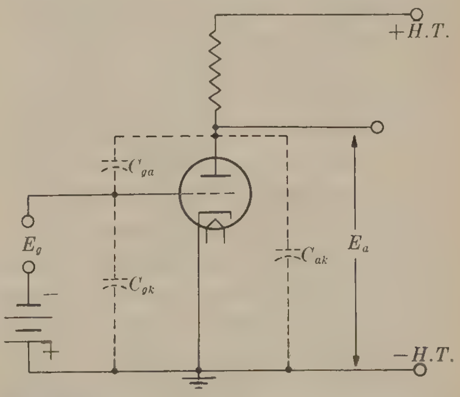


FIG.2-33 Interelectrode capacitances in a triode amplifier.

a-c ground are effectively in parallel and therefore add together. In addition C_{ga} depends upon the amplification of the valve, C_{ga} being greater when the gain is greater. The exact relationship between these capacitances, which is developed in the applied mathematics section at the end of this chapter, is given by the formula

$$C_i = C_{gk} + C_{ga}(1+A)$$

where C_i = input capacitance of the valve
C_{gk} = grid-to-cathode capacitance
C_{ga} = grid-to-anode capacitance
A = stage gain.

The input capacitance of the valve has a shunting or short-circuiting effect on the signal voltage which is applied between grid and ground. The shunting effect becomes more pronounced as the frequency of the signal is increased. This causes the gain of the circuit to fall off at high frequencies.

The output capacitance C_0 of the stage consists of the anode-to-cathode interelectrode capacitance C_{ak} in parallel with the stray capacitance between the anode wiring and ground. These capacitances are so small that their effect is negligible at low frequencies. At high frequencies, the reactance of C_o is considerably less. It presents a shunting effect in parallel with the load impedance which reduces the gain considerably. The frequency at which the response begins to fall off is determined to a large extent by the output capacitance of the valve. In order to extend the response of the amplifier toward the higher frequencies, it is necessary to maintain a small value of output capacitance. The equivalent circuit of Fig. 2-34 shows those components which are effective in determining the high-frequency response of the amplifier.

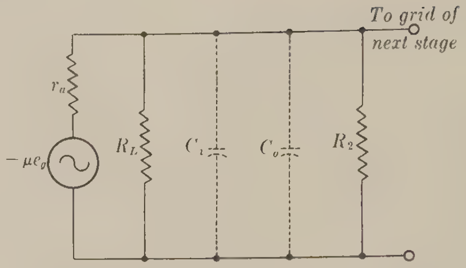


Fig. 2-34 Equivalent circuit of an a-f amplifier for high frequencies.

The frequency response of an amplifier is often extended by reducing the value of the anode load resistance, as illustrated in Fig. 2-35. The flat portion of the response curve is extended from 10,000 c/s to 50,000 c/s by reducing the load resistance from 500,000 ohms to 50,000 ohms in a particular amplifier. By this means, the output impedance is kept substantially greater than the load resistance, even at the higher frequencies. It should be noted, however, that this reduction in load resistance also causes the gain to be reduced to half its original value.

2.8 Pentode amplifiers. Pentodes are frequently used in resistance-capacitance coupled a-f voltage amplifiers. Pentode amplifiers provide a higher gain per stage for the same H.T. than triode amplifiers. A typical amplifier circuit employing a pentode valve is shown in Fig. 2-36. C_1R_1 make up the input coupling circuit and C_2R_2 provide cathode bias. R_4 is the anode load resistor and C_4 the output coupling capacitor. The suppressor grid is maintained at the same d-c poten-

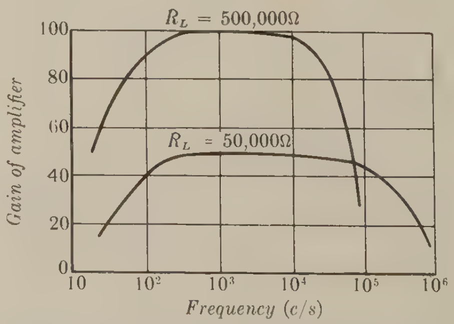


Fig. 2-35 Effect of reducing load resistance on the frequency-response curve of an amplifier.

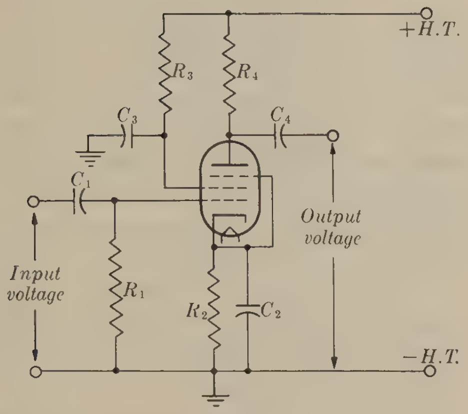


FIG. 2-36 Resistance-capacitance coupled pentode amplifier.

tial as the cathode and at a-c ground potential by C_2. R_3 and C_3 provide the correct a-c and d-c potentials to the screen grid. R_3 is the screen voltage dropping resistor. Screen current passing through this resistor drops the screen voltage from +H.T. to a value which allows the correct amount of space current to flow in the valve. The screen current contains alternating components similar to those in the anode current. If these components are allowed to pass through R_3, degeneration takes place. That is, if the control grid is made less negative, the anode and screen currents increase. The increase in screen current causes a larger p.d. across R_3, the screen voltage is reduced and the anode current is reduced. The reduction in anode current reduces the output voltage, decreasing the gain of the stage. To prevent degeneration, the by-pass capacitor C_3 is included in the circuit. This capacitor maintains the screen at a steady potential in the same manner as C_2 maintains the cathode at a steady potential. Since the cathode, screen and suppressor are kept at steady potentials, only the control grid and anode vary in potential. Hence the equivalent circuit of a pentode is the same as that of a triode.

The grid-to-anode capacitance of a pentode is much smaller than that of a triode and therefore the impedance between these two electrodes is much higher. In this way the screen grid effectively isolates the input from the output circuit. The input capacitance of a pentode is also much smaller than that of a triode. Thus a pentode amplifier has a smaller shunting effect on the output of the preceding stage.

Because of the reduced input capacitance and high amplification factor, the effect of stray wiring capacitance is more noticeable than with a triode valve. For this reason the circuits must be carefully laid out and may even require shielding.

The a-c resistance of a pentode is high, sometimes as much as one or two megohms. This means that changes in load impedance have a greater effect on the gain of a pentode stage than a triode stage. The amplification factor of a pentode is also high, in the neighbourhood of 1000. However, the gain is limited by the size of the load resistor. If it were made equal to the anode resistance, an excessively large H.T. voltage would be required to provide a reasonable anode potential. Regardless of this limitation, higher values of stage gain are obtained with pentode amplifiers than with triodes.

2.9 Effect of coupling components on formula for stage gain. In order to calculate the gain of an amplifier, it is necessary to refer again to the equivalent circuit diagrams.

A. Triode Amplifiers

It is pointed out in section 2.4 that the triode amplifier may be considered for purposes of analysis as being equivalent to a generator providing an a-c voltage of magnitude μe_g in series with the anode resistance of the valve r_a and the load resistance R_L as shown in Fig. 2-37. This type of equivalent circuit is generally

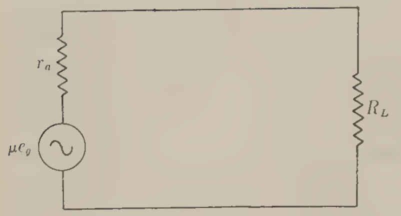


FIG.2-37 Equivalent circuit of triode amplifier.

referred to as a *constant-voltage-generator* form of equivalent circuit. The a-c component of anode current is found by applying Ohm's Law to the equivalent circuit and is given by the formula

$$i_a = \frac{\mu e_g}{r_a + R_L}$$

The a-c component of anode voltage, therefore, is

$$e_a = \frac{\mu e_g R_L}{r_a + R_L}$$

The amplification or gain of the amplifier is the ratio of anode voltage to grid voltage. That is

$$A = \frac{e_a}{e_g} = \frac{\mu R_L}{r_a + R_L}$$

or

$$A = \frac{\mu}{1 + \frac{r_a}{R_L}}$$

This formula may be expanded to calculate the gain of the complete circuit including the coupling capacitance and the grid-leak resistance of the following stage. For the mid-frequencies, the coupling capacitance is effectively a short circuit and may be omitted as shown in the equivalent circuit of Fig. 2-31. The effective load resistance R_L is then the anode resistance R_1 and the grid-leak resistance R_2 in parallel (Fig. 2-22). These resistances may be combined by

$$\frac{1}{R_L} = \frac{1}{R_1} + \frac{1}{R_2}$$

or

$$R_L = \frac{R_1 R_2}{R_1 + R_2}$$

Substituting for R_L in the gain formula,

$$A = \frac{\mu}{1 + \frac{r_a(R_1 + R_2)}{R_1 R_2}}$$

B. Pentode Amplifiers

In the case of a pentode amplifier in which the load resistance is a small fraction of the anode resistance of the valve, the gain is more dependent on the transconductance of the valve than on the amplification factor.

Substituting

$\mu = g_m r_a$ in the above formulas,

$$i_a = \frac{g_m e_g r_a}{r_a + R_L}$$

and

$$e_a = \frac{g_m e_g r_a R_L}{r_a + R_L}$$

From this formula, it appears that the amplifier is equivalent to a generator producing a current equal to $g_m e_g$ which is made to flow through r_a and R_L in parallel. The circuit therefore takes the form shown in Fig. 2-38 and is referred to as the *constant-current-generator* form of equivalent circuit. This form is more useful than the *constant-voltage-generator* form for the study of amplifiers containing tetrodes, pentodes and beam power valves.

The gain of the pentode amplifier is found as before by dividing the anode voltage by the grid voltage.

$$A = \frac{e_a}{e_g} = \frac{g_m r_a R_L}{r_a + R_L}$$

or

$$A = \frac{g_m r_a}{1 + \frac{r_a}{R_L}}$$

To obtain the gain of the entire circuit operating in the mid-frequency range, it

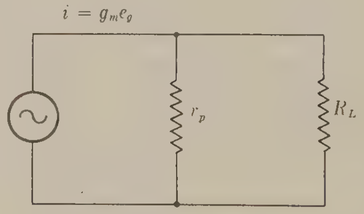


FIG.2-38 Constant-current-generator form of equivalent circuit.

is necessary to substitute

$$R_L = \frac{R_1R_2}{R_1+R_2}$$

$$\therefore A = \frac{g_m r_a}{1+\frac{r_a(R_1+R_2)}{R_1R_2}}$$

2.10 Distortion. In the preceding sections, it has been assumed that the waveform of the output voltage of an amplifier is the same as the waveform of the input voltage in every respect except amplitude. In practice, this condition is frequently difficult to achieve. Any change in the output waveform introduced by the valve or the circuit is called *distortion*. It is usually desirable to reduce distortion to a minimum in audio-frequency amplifiers as it causes voice and music to sound unnatural or even unintelligible. In video amplification, distortion may cause the observed patterns to be so blurred as to be useless. However, distortion is not always undesirable. Some radio and radar circuits such as detectors, frequency multipliers, differentiating circuits and other wave shapers make use of distortion.

The three main types of distortion with some of the more common sources of each are discussed in the following paragraphs.

A. Amplitude Distortion

Distortionless amplification can take place only when the valve is made to operate over the linear portion of the dynamic $e_c - i_b$ characteristic. Although no part of the mutual characteristic is strictly linear, the centre portion is sufficiently linear to give negligible distortion. If the valve is properly biased and the grid-signal amplitude restricted sufficiently to maintain operation in the linear region, the output waveform is an exact amplified copy of the input waveform.

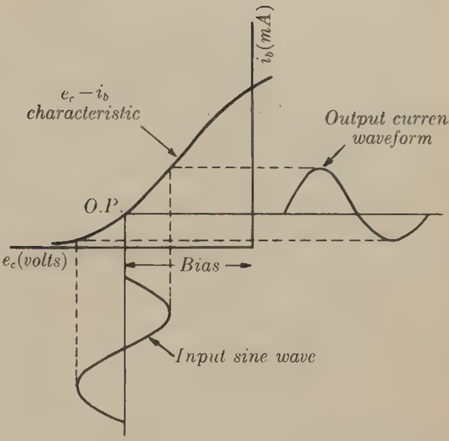


FIG. 2-39 Operation on the lower-bend portion of the characteristic.

If, however, the valve is made to operate over the curved portion of the characteristic, either by the use of incorrect bias or by applying too large a signal voltage to the grid, amplitude distortion results. In this case the amount of gain changes as the amplitude of the signal changes. One type of amplitude distortion is illustrated by the diagram of Fig. 2-39. Too much negative bias is applied to the grid of the valve so that it operates over the lower curved portion of the characteristic. This causes the anode current to increase more on the positive half cycle than it decreases on the negative half cycle. Hence one half of the waveform is amplified more than the other half and distortion is introduced. The average anode current is greater when the signal is applied than for no signal. This type of distortion may be detected by inserting a d-c milliammeter in the anode circuit. If the anode current increases when a signal is applied to the grid, then too much negative grid bias has been used.

If the grid bias is too small, the valve operates on the upper bend of the characteristic as shown in Fig. 2-40. In this case the negative half cycle of the signal is amplified more than the positive half. The average anode current is less than the no-signal value and a milliammeter inserted in the anode circuit registers a

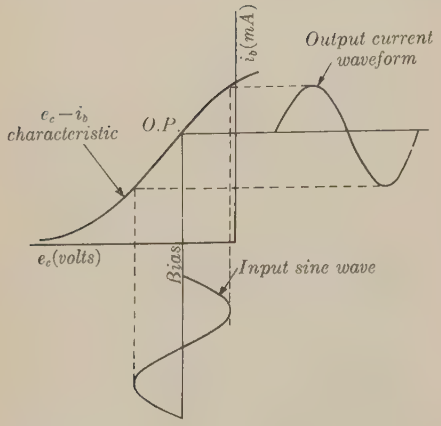


FIG.2-40 Operation on the upper-bend portion of the characteristic.

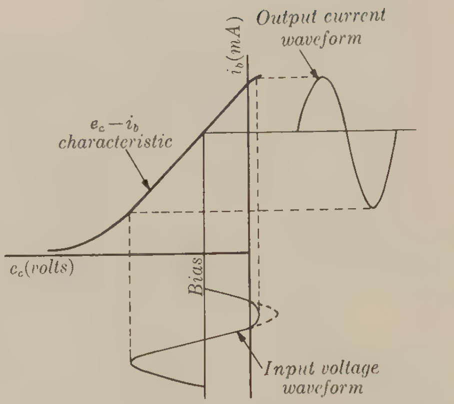


FIG.2-42 Distortion due to grid current flow.

decrease in current when a signal is applied to the grid.

An amplifier stage may be operated with such a bias and input that the positive half cycles of grid-signal voltage are large enough to drive the grid positive for part of each cycle. The resulting distortion may be understood by referring to the diagram of Fig. 2-41. The input voltage E_i applied across C_c and R_g causes a current I_i to flow through R_g. During the positive half cycle of the input voltage, electrons flow from ground through R_g to charge C_c. It is this flow of electrons through R_g which maintains the grid-cathode voltage. If the grid is driven positive to the cathode, electrons are attracted to the grid to charge C_c. Hence the voltage across R_g is reduced and the grid voltage does not rise as high as the signal voltage rises. The greater the grid current flow, the smaller the effective positive half cycle of the grid-signal voltage. Hence the increase in anode current is not as great as the decrease on the negative half cycle when no grid current flows. The resulting distortion is represented graphically in Fig. 2-42. The dotted portions of the input and output curves represent the waveforms which would exist if no grid current were to flow. The remedy for these types of distortion is to increase the anode voltage so that a longer straighter portion of the characteristic is obtained, or to select a valve which can handle the input without producing distortion. The following examples illustrate the operation ranges which permit distortionless amplification for two triode valves of different characteristics. The dynamic mutual characteristics for the two valves are illustrated in Fig. 2-43.

For Valve 1 the grid-swing range for distortionless amplification is 18V. For Valve 2 the range is 32V. Valve 1 is therefore able to handle a grid signal of 9V amplitude while Valve 2 can handle an amplitude of 16V.

It should be noted that Valve 1 has higher values of amplification factor and anode resistance. Hence the following

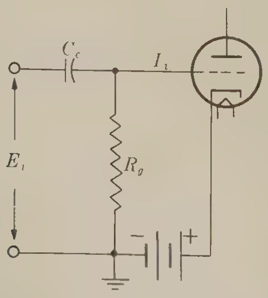


FIG.2-41 Grid current flow.

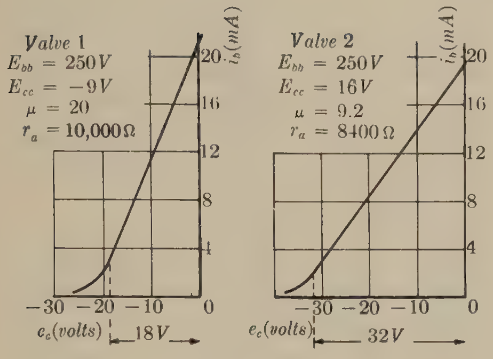


Fig.2-43 Operating region for distortionless amplification.

general rule is indicated. The greater the grid swing to be handled by a valve, the lower should be the amplification factor and a-c resistance of the valve if the output is to be undistorted.

The effect of amplitude distortion is the generation of new frequencies, either harmonics or sum and difference frequencies. For this reason, amplitude distortion is sometimes called *harmonic distortion.*

Amplitude distortion may be introduced in other parts of the circuit than the valve itself. Coupling devices, especially transformers, may have non-linear characteristics similar to those of valves. Also, audio devices such as earphones, speakers and microphones may introduce distortion. Weak magnets, and magnetic saturation due to excessive currents, frequently produce this effect.

Amplitude distortion is not always undesirable, however. It is utilized in squaring or limiting circuits, detectors and frequency multipliers.

B. Frequency Distortion

It is pointed out in section 2.8 that it is difficult to design a circuit which will provide uniform amplifications of voltages over a wide range of frequencies. An audio-frequency amplifier is constructed to amplify as uniformly as possible voltages at frequencies contained in normal speech and music. A complete picture of the gain of the circuit is provided by the frequency-response curve. Uniform amplification takes place over the range of frequencies for which the response curve is flat. The amplification normally falls off at the very low and very high frequencies. This is indicated by a drop in the response curve over these frequencies. The reduced gain at the extreme frequencies is caused by the change in reactances of coupling capacitors, inductive loads, self-capacitances of windings, interelectrode capacitances and other non-resistive components.

If the waveform of the input voltage to an amplifier is not a sine wave, the voltage contains components at various frequencies. If some of these components are amplified more than others, the waveform of the output voltage is different from that of the input and distortion has been introduced. This type of distortion is called *frequency distortion* since it is caused by a variation in voltage gain at various frequencies.

Frequency distortion is also used to advantage in certain circuits such as selective amplifiers, differentiating circuits, frequency multipliers, filters and tone controls.

C. Phase Distortion

The output voltage of an amplifier with a resistance load is 180° out of phase with the input voltage. If the load is reactive, this phase difference between output and input voltages is not 180°. Moreover it varies with changes in frequency. Hence, if the input voltage contains components of various frequencies, the phase relationships of the components are changed in the output. As a result the shape of the output waveform differs from that of the input and distortion has been introduced. Phase distortion usually occurs in the circuits with reactive components such as coupling circuits, grid bias circuits and decoupling circuits.

Phase distortion is negligible in most a-f amplifiers. In video amplifiers however, where uniform amplification is required over an extremely wide range of frequencies, phase distortion is one of the chief types and one of the most difficult to eliminate.

In circuits used as wave shapers, phase distortion as well as frequency distortion is frequently utilized to change the shape of the waveform.

2.11 Audio-frequency power amplifiers. The purpose of the amplifiers discussed in the previous sections is to provide an alternating output voltage which is an amplified replica of the input voltage. The prime consideration in the design of the voltage amplifier is to obtain the highest possible gain provided that distortion is not introduced in any large degree.

The power amplifier on the other hand is designed to deliver maximum undistorted power to a load impedance while the voltage gain is of minor importance. A typical a-f power amplifier used to drive a loudspeaker is shown in Fig. 2-44. The circuit is similar in construction and operation to the voltage amplifier. The chief differences between the two types of amplifiers lie in the design of the valves and in the type of load impedance used.

The valve is biased so that the operating point is at the middle of the straight portion of the $e_c - i_b$ characteristic. Bias is provided by the resistor R_2 and the a-f by-pass capacitor C_2 in the cathode circuit. The values of these components are selected from the same considerations as those which apply to voltage amplifiers. The a-f signal voltage is coupled to the grid of the amplifier through the coupling capacitor C_1. The grid-leak resistor R_1 provides a ground return for any electrons which might accumulate on the grid.

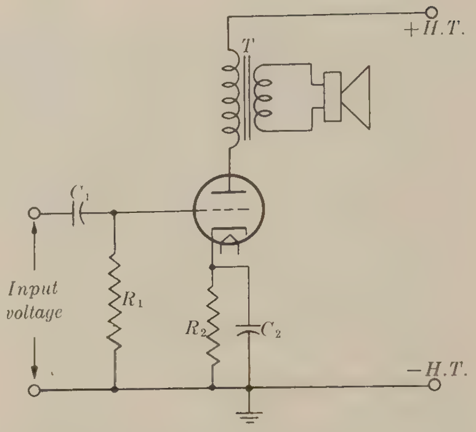


FIG.2-44 Audio-frequency power amplifier.

The anode load impedance is the primary circuit of the output transformer *T*. The variations in the anode current caused by the grid-signal voltage induce an a-c current in the transformer secondary and this current flows through the speaker. Since the speaker is a current-operated device, it is desirable to cause as large a current as possible to flow through it. This is obtained by the use of a valve which conducts a very large anode current and by the use of the transformer. The transformer provides a step down in voltage and therefore a step up in current from the anode circuit to the speaker. In terms of impedances, the transformer is used to match the higher impedance of the anode circuit to the lower impedance of the speaker coil.

Power valves, whether used in receivers or in transmitters, must be capable of passing a heavy current to the load impedance. Consequently the power valve is fitted with a large cathode which is capable of high electron emission. Large anodes are also used to dissipate the heat produced within the valve. Since the power valve is preceded by one or more stages of a-f voltage amplification, it must be designed so that it can handle large input voltages without overloading. This requirement has led to the development of valves having a mutual characteristic which is linear for comparatively wide swings of grid voltage.

A. Maximum Power Output

The primary consideration in a power amplifier is to secure maximum power output within the permissible limits of distortion. A fundamental law of electricity is that maximum transfer of energy from a source to a load takes place when the impedance of the load is equal to the impedance of the source. This law is true whether the source of power is a battery, a generator, or a valve supplying power to a load. In the case of the power amplifier circuit, therefore, maximum power is transferred from the valve to the speaker when the impedance of the load circuit is made equal to the anode resistance of the valve. Expressed in mathematical symbols, Z_L must equal r_a.

The power relations in a triode amplifier with a load resistance R_L are illustrated by means of the dynamic $e_b - i_b$ characteristics of the valve as shown in Fig. 2-45. The steady operating anode voltage is E_b and the corresponding steady anode current is I_b. Thus point P on the load line represents the operating point. The grid-signal voltage causes the anode current to vary from i_b minimum to i_b maximum. The corresponding limits of the anode voltage are e_b maximum and e_b minimum. The useful power output is the a-c power developed in the load and is given by the formula

$$P_{a\text{-}c} = E_a I_a$$

where $P_{a\text{-}c}$ = a-c power output in watts

E_a = effective value of the a-c component of anode voltage

I_a = effective value of the a-c component of anode current.

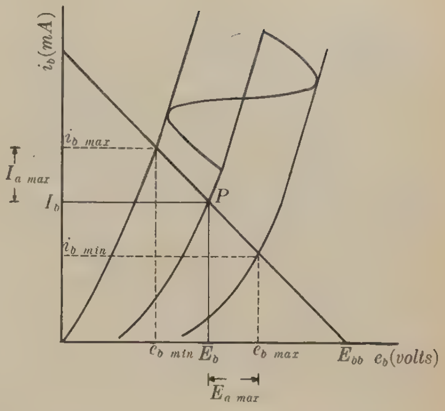


FIG.2-45 Operation of a-f power amplifier.

The peak values of a-c anode voltage and current are observed from Fig. 2-45 to be E_a maximum and I_a maximum respectively,

where $E_{a\ max} = e_{b\ max} - E_b$

and $I_{a\ max} = i_{b\ max} - I_b$.

But the effective value of a sine-wave voltage or current is equal to the peak value divided by $\sqrt{2}$.

i.e. $$E_a = \frac{E_{a\ max}}{\sqrt{2}}$$

and $$I_a = \frac{I_{a\ max}}{\sqrt{2}}$$

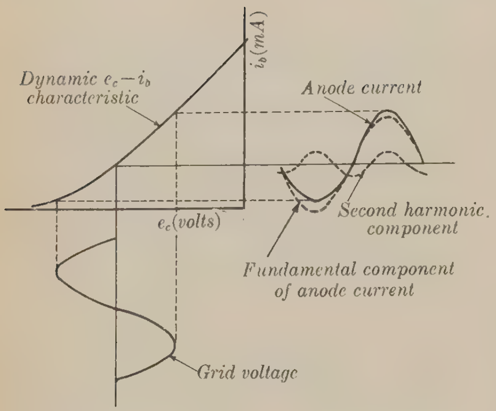
Consequently $$P_{a\text{-}c} = E_a I_a = \frac{E_{a\ max}}{\sqrt{2}} \times \frac{I_{a\ max}}{\sqrt{2}}$$

$$\therefore P_{a\text{-}c} = \frac{E_{a\ max} \times I_{a\ max}}{2}$$

That is, the useful output power in the anode circuit equals half the product of the peak values of a-c anode voltage and current.

B. Distortion In Power Amplifiers

Since the anode resistance of a triode is small, the load impedance must also be small to obtain maximum power output. When a valve is operated with such a low value of load impedance, a considerable amount of distortion is introduced. The distortion is due to the curvature of the dynamic $e_c - i_b$ characteristic for low values of R_L. As a result, the positive half cycle of the grid-signal



$i_b(mA)$
Dynamic $e_c - i_b$ characteristic
Anode current
e_c(volts)
Second harmonic component
Fundamental component of anode current
Grid voltage

FIG. 2-46 Distortion in a power amplifier.

voltage is amplified more than the negative half. This effect is illustrated in Fig. 2-46, in which a sine waveform of grid voltage produces an asymmetrical waveform of anode current. If the anode current waveform is resolved into its sinusoidal components, it is found to consist chiefly of the fundamental and second harmonic components as shown by the dotted waveforms. The ratio of the peak values of second harmonic and fundamental components expressed as a percentage is called the *percentage distortion.* Distortion in the amplifier may be reduced by increasing the value of the anode load resistance as this tends to reduce the curvature of the characteristic.

The a-c power output, when operation is restricted to the linear portion of the characteristic, is given by the formula

$$P_{a\text{-}c} = I_a^2 R_L$$

where I_a = effective value of the a-c component of anode current

R_L = effective resistance of the load.

Also, $$I_a = \frac{\mu E_g}{r_a + R_L}$$

as derived from the equivalent circuit.

If the load is pure resistance, the a-c power output is a maximum when

$$R_L = r_a$$

$$\therefore I_a = \frac{\mu E_g}{2r_a}$$

The maximum a-c power output is then given by

$$P_{a\text{-}c\ max} = \frac{\mu^2 E_g^2}{4r_a^2} \times r_a$$

$$= \frac{\mu^2 E_g^2}{4r_a}$$

$$\because \qquad \mu = g_m r_a$$

$$\therefore P_{a\text{-}c\ max} = \frac{E_g^2}{4} \mu g_m$$

That is, the maximum a-c power output is proportional to the square of the grid-signal voltage and the factor μg_m which depends only on the construction of the valve. This maximum a-c power output is accompanied by considerable distortion.

C. Optimum Power Output

The optimum power output is a compromise between the maximum power output obtained when $R_L = r_a$, and the power output which may be obtained with a reasonable amount of distortion. The optimum power output is obtained when the load impedance is approximately equal to twice the anode resistance of the valve. In this case there is less than 5% distortion and the power output is only about 10% lower than the maximum value. Thus, in a power amplifier circuit, the choice of load impedance is determined by the value of r_a in the valve selected, whereas in the voltage amplifier circuit the load impedance should be as large as possible to obtain maximum gain.

D. Power-Amplification Ratio and Power Sensitivity

In a power amplifier the term corresponding to the gain of a voltage amplifier is the power-amplification ratio. It is the ratio of the a-c power output to the a-c power consumed in the grid circuit. If the amplifier is operated so that the grid signal does not exceed the grid bias, grid current does not flow. Hence the power supplied to the grid circuit is negligible and the power amplification is extremely high. In this case the term may have little significance and the performance of the amplifier may be better represented by its power sensitivity which is defined by the relationship

$$\textit{Power Sensitivity} = \frac{P_{a\text{-}c}}{E_g^{\,2}}$$

and is expressed in mhos.

This ratio provides a convenient measure for comparing power valves since it expresses the relationship between power output and the amount of input voltage required to produce that power.

E. Anode Efficiency and Anode Dissipation.

In a power amplifier, d-c power is supplied to the valve through the anode circuit by the H.T. supply. The useful power output is a-c power. Thus a power amplifier converts d-c power to a-c power under the control of the alternating voltage applied to the grid of the valve.

Some of the input power is lost in the valve as heat and some appears in the load as d-c power in addition to the useful a-c power output. The ratio of a-c power developed in the load to the d-c power input to the valve expressed as a percentage is called the *efficiency* of the anode circuit. The anode efficiency is increased by reducing the d-c component of anode current or by increasing the a-c component.

In power amplifiers in which the valve is biased so that the operating point is at the centre of the straight portion of the characteristic and distortion is low, the anode efficiency seldom exceeds 25%. Since the power output is proportional to the square of the alternating component of the anode current, a small a-c component of anode current results in a low power output in comparison with the power input and hence low efficiency.

The amount of power lost in the valve as heat is called the *anode dissipation.* It is equal to the difference between the d-c power input to the valve and the a-c power output. In valves in which the grid is biased half way to cut-off, the power input to the valve remains constant and is independent of input voltage variations over the linear portion of the characteristic. However, the a-c power output varies as the square of the input voltage for a given load resistance. Thus the a-c power output increases and the anode dissipation decreases when the grid-signal voltage increases. That is, the valve operates at a lower temperature when delivering a large amount of power to the load than when producing a small output or none at all.

The relationships between the d-c power input to the valve, the anode dissipation and the a-c power output in this type of amplifier are shown graphically in Fig. 2-47 for one cycle of sine-wave voltage applied to the grid. The d-c and a-c grid voltages are shown in Fig. 2-47 (a). The anode current and anode voltage waveforms are shown in Fig. 2-47 (b) and (c). The energy supplied to any part of the circuit during one cycle of operation equals the product of the power in that component and the duration of the cycle. The total energy

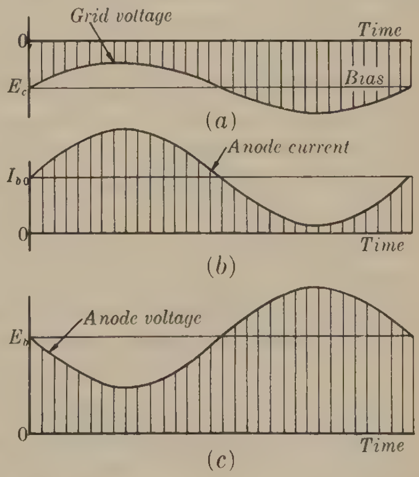


FIG.2-47 Anode efficiency of a power amplifier.

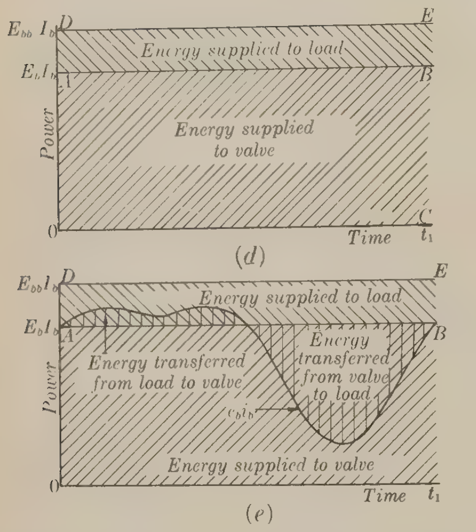


FIG. 2-47 Anode efficiency of a power amplifier (d) Distribution of energy with no grid voltage, (e) Distribution of energy with grid voltage.

supplied by the power supply during one cycle is made up of two parts, the energy supplied to the valve and the energy supplied to the load. The total energy equals the product of the supply voltage E_{bb}, the anode current I_b and the duration t_1, and is represented by the area *ODEC*, Fig. 2-47 (d). When no signal is applied to the grid, the energy supplied to the valve equals the product of the anode voltage E_b, the anode current I_b and the time t_1. This is represented by the area *OABC*. Hence the area *ADEB* represents the energy supplied to the load.

The distribution of energy between the valve and the load when the signal voltage is applied is shown by the graph of Fig. 2-47 (e). The instantaneous power in the valve equals e_bi_b. Hence the energy supplied to the valve during one cycle is represented by the area under the e_bi_b curve. The area above this curve represents the energy supplied to the load. The graph indicates that there is a transfer of energy from the load to the valve during the first half of the cycle and from the valve to the load during the second half cycle. The a-c energy output from the valve to the load is equal to the difference between the heavily shaded area above *AB* and the heavily shaded area below *AB* and equals approximately 25% of the energy supplied to the valve. Hence the anode efficiency of the valve is approximately 25%.

The maximum anode-dissipation rating of a valve is the maximum amount of power which it is able to absorb and radiate or conduct away safely as heat. It depends upon the maximum temperature that the anode can reach without giving off absorbed gas or emitting electrons. It sometimes happens that the input voltage is accidentally removed from a valve which is operating close to its maximum anode dissipation with the result that the temperature is suddenly increased above the allowable value and the valve is destroyed.

F. Parallel Operation.

A power output greater than that which a single valve provides is obtained by connecting two or more valves in parallel. Fig. 2-48 shows the method of connecting two valves, usually of the same type, in parallel in an a-f power amplifier. The two grids are connected together and the signal voltage applied

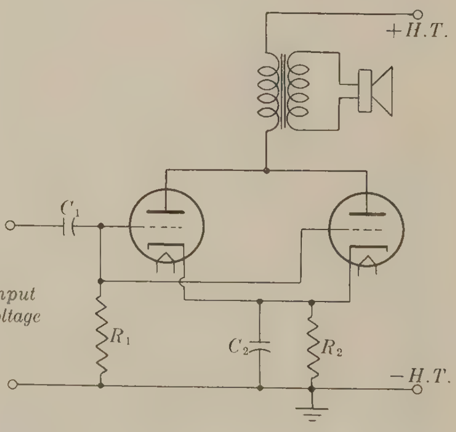


FIG. 2-48 Operation of valves in parallel.

to both. The two cathodes are connected together and a common bias network consisting of R_2C_2 serves the two valves. The anodes are also connected together and the two anode currents flow through the primary winding of the output transformer. The input voltage which can be handled without overloading is no greater than that for a single valve but the total anode current is double that of either valve working alone. The effective anode resistance of the two valves is therefore half the resistance of a single valve. The output transformer must match this reduced impedance to the impedance of the load circuit.

2.12 Push-pull amplifier. The distortion of the output voltage waveform due to the curvature of the mutual characteristic is greatly reduced by connecting two valves in a push-pull circuit. This reduction in distortion makes available an optimum power output which is more nearly equal to the maximum power output obtainable from the valves.

A push-pull circuit using transformer input coupling and transformer output coupling is shown in Fig. 2-49. The circuit is a typical a-f power amplifier used to furnish undistorted power to a loudspeaker. The two valves are of the same type and while triodes are shown in the diagram, beam power valves are also commonly employed in push-pull circuits. The signal voltage is applied through the input transformer T_1 to the grids of V_1 and V_2. This transformer has a centre-tapped secondary winding. The centre tap is maintained at a-c ground potential by the low internal resistance of the bias battery C. This battery also keeps the grids of V_1 and V_2 somewhat negative to their cathodes so that the signal voltage does not drive the grids positive. The anode voltages for the two valves are supplied from the positive terminal of the H.T. supply through the primary winding of the output transformer T_2.

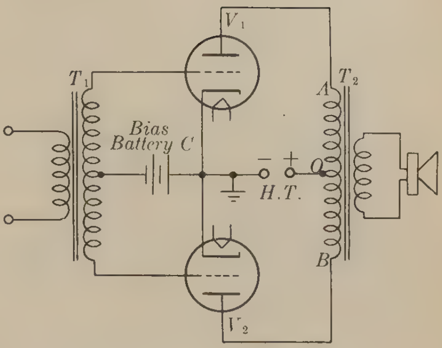


FIG.2-49 Push-pull amplifier.

It should be noted that the two valve circuits are exactly similar in every respect and that each contains all the components of a single-valve a-f power amplifier.

The signal voltage applied to the primary winding of T_1 causes a voltage to be induced in the secondary winding. Since the centre tap is maintained at a-c ground potential, the a-c voltages applied to the grids of V_1 and V_2 are 180° out of phase with each other. That is, as the grid of V_1 is made less negative, the grid of V_2 is made more negative, as shown on the dynamic $e_c - i_b$ curves of Fig. 2-50. As a result, the current through V_1 increases while the current through V_2 decreases. This causes an increasing electron flow in the direction AO and a decreasing electron flow B to O in the primary winding of the output transformer. The currents induced in the secondary winding of T_1 by these primary currents are in phase with each other and the resultant current is equal to their sum.

During the next half cycle of signal voltage, the grid of V_2 is made less negative and the grid of V_1 is made more negative. Hence the electron flow through V_2 and the primary of T_2 increases in the direction BO, while the electron flow through V_1 decreases in the direction AO. The secondary currents are again in phase with each other and

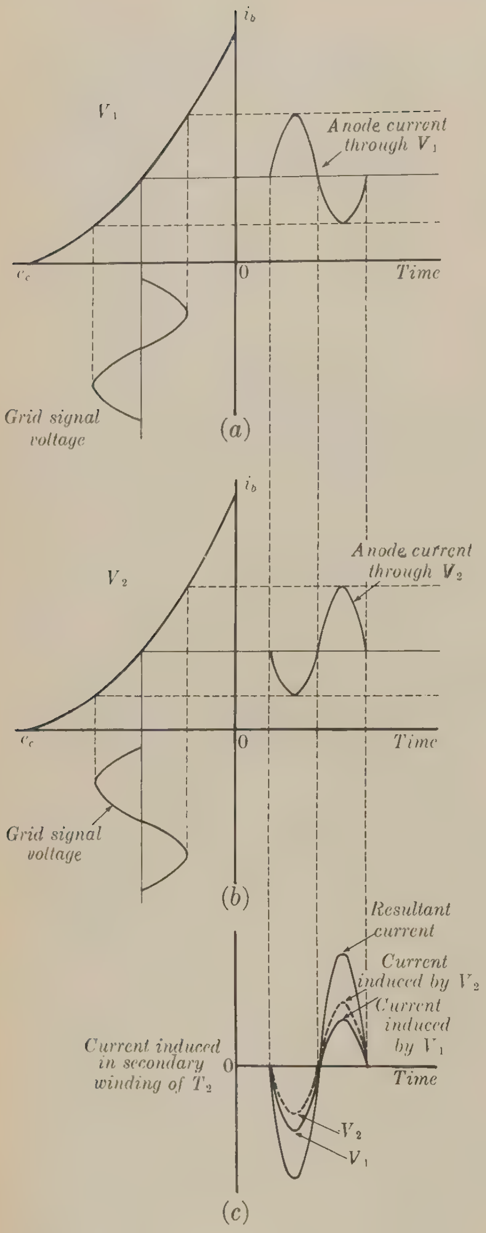


FIG.2-50 Operation of a push-pull amplifier.

flow in the opposite direction to the resultant current during the previous half cycle. The waveforms of Fig. 2-50 show that the sinusoidal input voltage is distorted in the output of each valve due to operation over the curved portion of the characteristic. However, the large positive half cycle of current through V_1 occurs at the same time as the small negative half cycle of current through V_2 and vice versa. Therefore the same amplitude of current flows in the secondary winding of T_2 during both half cycles and the distortions cancel each other.

The resultant waveform of current through the secondary of T_2 is slightly flattened at the peaks. This type of distortion is different from that present in a single-valve amplifier in that the even harmonics cancel each other. Only the odd harmonics such as the third, fifth, seventh, etc., are introduced and so the amount of distortion is smaller than that produced by a single valve or by two valves in parallel producing the same amount of output power. Alternatively, the push-pull circuit may be made to deliver up to 50% more power than the parallel circuit with no increase in distortion. This increase is effected by changes in the operating voltages and the effective load resistance which are permissible in a push-pull circuit. Moreover, since distortion is reduced by the cancellation of the even harmonics, the region of operation of the valve may be extended farther into the curved portion at the lower end of the characteristic curve. That is, a greater bias and a greater amplitude of voltage may be applied to the grids without exceeding the allowable percentage of distortion. In this way, a larger power output is obtained and the efficiency of the circuit is increased.

Three classes of operation of the valves are in common use in a-f push-pull amplifiers. They are Class A, Class AB and Class B and are described in the following paragraphs.

A *Class A amplifier* is one in which the bias is adjusted so that the operating point is approximately midway between cathode potential and cut-off potential and in which the grid-signal voltage does not go beyond cut-off. Since the grid is not driven beyond cut-off, anode current flows through the valve during all parts of the cycle. However, in some cases of Class A operation the grid may be driven positive to the cathode and grid current may flow.

A *Class AB amplifier* is one in which the bias is increased so that the operating point is well down on the characteristic curve. On the negative half cycle the peak of the signal drives the grid beyond cut-off. Hence anode current flows for appreciably more than half but less than the entire alternating grid-voltage cycle. In the case where the grid signal voltage exceeds the bias, grid current flows for part of the cycle.

In a *Class B amplifier*, the grid bias is approximately equal to cut-off so that the anode current is almost zero when no input voltage is applied. Hence the anode current flows for approximately one-half of each cycle when an alternating grid voltage is applied.

Class B operation of valves in a push-pull circuit is the most efficient type of operation possible for a-f power amplifiers. The waveforms of Fig. 2-51 indicate the operation of the valves when the grid signal is applied. During the positive half cycle of grid voltage, one valve conducts current while the other is cut off. During the negative half cycle of the input voltage, the operation of the two valves is interchanged. The two valves supply power to the load alternately. The resultant waveform of the output voltage is distorted only slightly more than in the case of Class AB operation described in connection with Fig. 2-50. The distortion is maintained at a very low value if the dynamic $e_c - i_b$ characteristic is almost linear over the operating region. This type of operation provides a large amount of a-c power output with moderate-sized valves and with a relatively small supply of d-c power to the circuit.

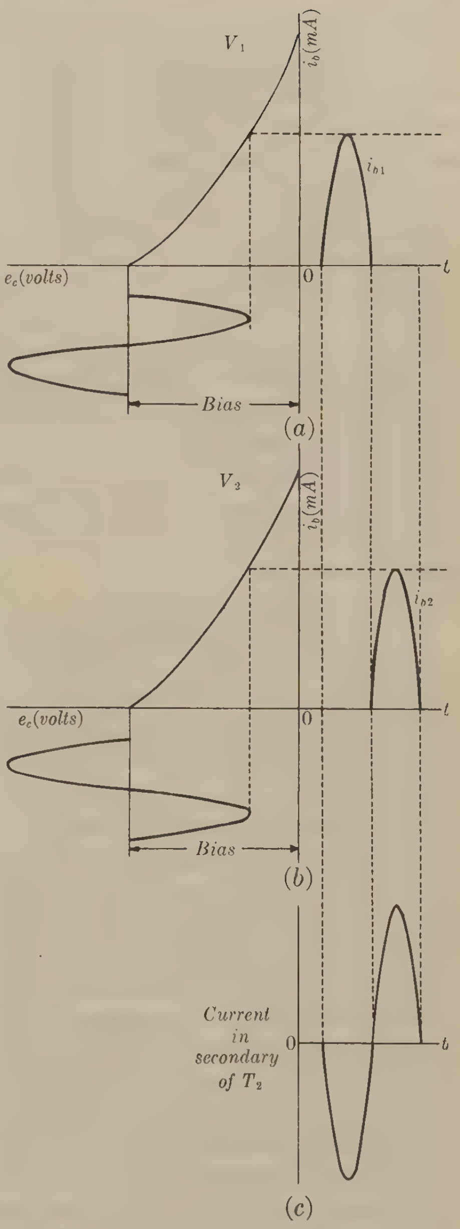


FIG. 2-51 Class B operation of a push-pull amplifier.

2.13 Radio-frequency voltage amplifiers. Radio-frequency amplifiers are employed in practically all radio receivers and transmitters. They are used in receivers to amplify the extremely small voltages induced in the aerial circuit by incoming radio waves. In transmitters they are used to increase the voltage and power level of the r-f carrier wave before passing it to the aerial circuit for radiation. Amplification at radio frequencies is the same in principle as a-f amplification and any differences in the valves and circuits employed are the result of differences in the frequency of the signal. R-f amplifiers are designed to amplify selectively at a single radio frequency or over a narrow band of frequencies. In order to satisfy this condition, r-f amplifiers commonly make use of resonant circuits tuned to the signal frequency in both the input and output circuits.

A. R-f Voltage Amplifier Circuit

A typical r-f voltage amplifier is shown in Fig. 2-52. A pentode valve is used as it provides higher gain and more stable operation than a triode at radio frequencies. The screen grid construction of the pentode reduces the amount of energy fed back from the output to the input circuit. This is required in order to prevent self-oscillation in the circuit. The tendency to oscillate becomes more pronounced as the frequency of the input signal is increased. R_2 is a screen voltage-dropping resistor and C_3 is the screen by-pass capacitor which maintains the screen at a-c ground potential. Bias for Class A operation of the valve is provided by R_1C_2 in the cathode circuit.

The signal to be amplified is coupled by the r-f transformer T_1 to the grid-cathode circuit of the valve. The secondary winding of this transformer and the variable capacitor C_1 constitute a resonant circuit which is tuned to the frequency of the signal. The input capacitance of the valve, which is chiefly grid-to-cathode capacitance in a pentode forms part of the effective tuning capacitance. In this way the shunting effect is made to serve a useful purpose and a high input impedance is maintained. The a-c signal applied to the grid of the valve causes the anode current to vary at the same frequency. The anode current flows through the load circuit which is also a resonant circuit tuned to the frequency of the signal. The output capacitance of the valve is used as part of the effective capacitance of this circuit. The d-c component of anode current flows through the inductor L. The d-c resistance of this path is very small so that the steady anode voltage is almost equal to the supply voltage. The a-c component of the anode current sets up oscillations in the load circuit at the frequency of the signal. The a-c voltage developed in the tuned circuit is coupled to the following circuit by C_5 and R_{g2}.

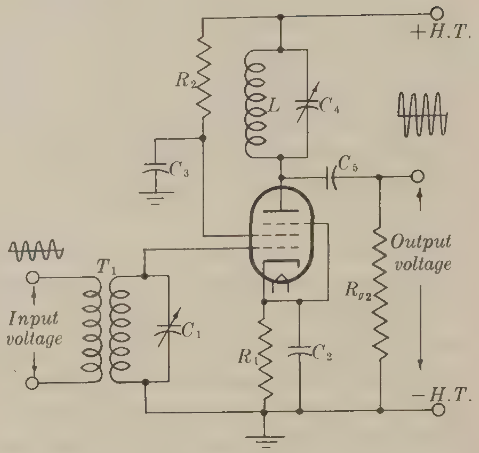


FIG.2-52 Radio-frequency voltage amplifier.

In order that the voltage gain at resonance be large, the resonant impedance of the output circuit LC_4 must be large in comparison with the a-c resistance of the valve. One method of increasing the impedance at resonance is to increase the $\frac{L}{C}$ ratio. However, in this case the selectivity of the amplifier is sacrificed and a compromise must be made between the two. A better method of increasing the resonant impedance and the voltage gain of the amplifier is to reduce the resistance of the resonant circuit to a minimum. This also results in an increase in selectivity.

In a tuned-anode amplifier the load impedance at resonance is equal to $\frac{L}{CR}$ and is resistive. Thus the formula for the gain of the amplifier developed in section 2.4 becomes

$$A = \frac{\mu \frac{L}{CR}}{r_a + \frac{L}{CR}}$$

B. *Equivalent Circuit of R-f Voltage Amplifier*

A formula for the gain of an r-f amplifier (Fig. 2-52) taking into account the grid-leak resistance R_{g2} of the following stage is developed from the equivalent circuit of Fig. 2-53. The pentode valve is represented as a source of constant current of value $i = g_m e_{g1}$ since the changes in anode voltage have practically no effect on the anode current over the range of normal operation. The anode resistance of the valve r_a is in parallel with the source of current as explained in section 2.9 B. The grid-leak resistance R_{g2} and the tuned circuit consisting of C, L and R are in parallel with r_a. The coupling capacitor is not shown in the equivalent circuit because its reactance is small in comparison with the resistance of R_{g2}. The interelectrode and stray capacitances in the anode circuit form part of the effective tuning capacitance C. The resistance R is partly the resistance accounting for power losses in the inductor and capacitor and partly the load in which the a-c power of the amplifier is dissipated. At the resonant frequency, the impedance of the tuned circuit is a resistance, $R_t = \frac{L}{CR}$.

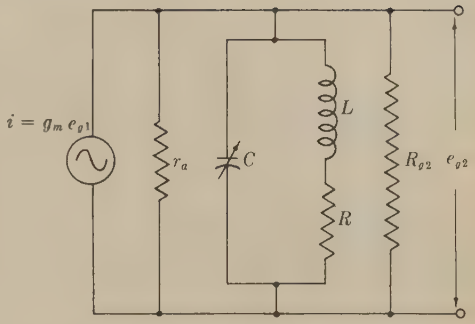


FIG.2-53 Equivalent circuit of r-f voltage amplifier.

The current from the source is, by Ohm's Law, equal to the output voltage e_{g2} divided by the effective resistance of r_a, R_{g2} and R_t in parallel.

That is $i = e_{g2}\left(\frac{1}{r_a}+\frac{1}{R_{g2}}+\frac{1}{R_t}\right)$

Also, $i = g_m e_{g1}$ where g_m is the mutual transconductance of the valve.

Therefore $g_m e_{g1} = e_{g2}\left(\frac{1}{r_a}+\frac{1}{R_{g2}}+\frac{1}{R_t}\right)$

Hence the voltage amplification,

$$A = \frac{e_{g2}}{e_{g1}} = \frac{g_m}{\frac{1}{r_a}+\frac{1}{R_{g2}}+\frac{1}{R_t}}$$

Thus the total impedance which affects the voltage gain of the amplifier is the parallel combination of r_a, R_{g2} and the tuned circuit. If the valve is a pentode with a high anode resistance the term $\frac{1}{r_a}$ is negligible and may be omitted from the formula.

C. *Single-Tuned Transformer Coupling*

An air-core transformer is often used for coupling in an r-f amplifier as shown in Fig. 2-54. The H.T. is fed through the primary of the transformer which provides the load in the anode circuit. The varying current through the primary winding induces an alternating e.m.f. across the secondary which forms part of the input circuit to the second valve. The transformer usually has a 1:1 turns ratio. Since the rotor plates of the variable capacitors in both tuned circuits

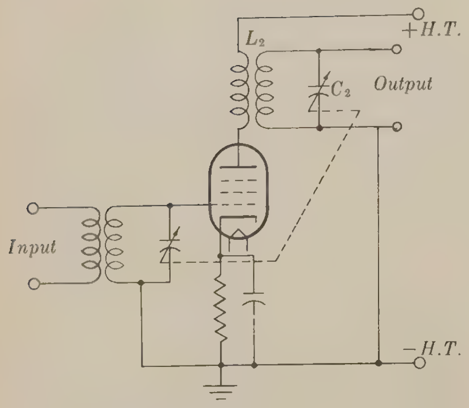


FIG. 2-54 Single-tuned transformer coupling.

are at ground potential, it is possible to combine the two capacitors into a single ganged unit. This permits simultaneous tuning of both circuits by a single control.

In order to prevent feedback of energy from the anode to grid circuit externally, careful layout of the wiring is necessary. The anode and grid leads must be kept short and separate from each other. The coils are usually enclosed in grounded shields to prevent radiation. By-pass capacitors must not only be of the proper type for efficient operation at the frequencies involved, but must be correctly placed with short leads. These precautions are necessary to guard against oscillation and degeneration.

The equivalent circuit and mathematical derivation of the gain of a transformer-coupled amplifier is somewhat more involved than in the case of the tuned-anode capacitance-coupled circuit of Fig. 2-52. The overall gain of the circuit depends upon the coupling between the primary and secondary circuits of the transformer. For optimum coupling

$$M = \frac{\sqrt{r_aR_2}}{2\pi f}$$

where m = mutual inductance
f = frequency of the signal voltage
r_a = a-c anode resistance of the valve
R_2 = effective resistance in the secondary circuit of the transformer.

Under these conditions, the gain of the circuit is given by the formula

$$A = \frac{\mu\omega L_2}{2\sqrt{r_aR_2}}$$

or since $\omega = \frac{1}{\sqrt{L_2C_2}}$ and $r_a = \frac{\mu}{g_m}$

$$A = \frac{1}{2}\sqrt{\frac{\mu g_m L_2}{R_2C_2}}$$

where L_2 and C_2 are the effective inductance and capacitance of the tuned secondary circuit of the transformer. However, when a valve with high anode resistance, such as a tetrode or pentode, is used the transformer coupling is much smaller than the optimum value. In this case the gain is more closely given as

$$A = \frac{\mu\omega^2 M L_2}{r_aR_2}$$

where A = stage gain
μ = valve amplification factor
M = mutual inductance
L_2 = effective secondary inductance
R_a = anode resistance of valve
R_2 = effective secondary resistance.

Thus, in order that the voltage gain be large the parallel tuned impedance of the secondary circuit and the quantity μg_m, which depends only on the valve, must both be large.

D. *Band-pass Coupling*

In some applications where the circuit is required to amplify voltages over a wide band of frequencies, tuning capacitors are connected across both the primary and secondary windings of the coupling transformer. This is referred to as *double-tuned transformer coupling* or *band-pass coupling*, (Fig. 2-55). When the proper degree of coupling is employed, uniform amplification is provided by the circuit over a band of frequencies with very little gain outside these limits. This is indicated by a frequency response curve with steep sides and a flat top (Fig. 2-56).

Amplifiers of this type are used in the intermediate-frequency (i-f) stages of superheterodyne receivers. Since the

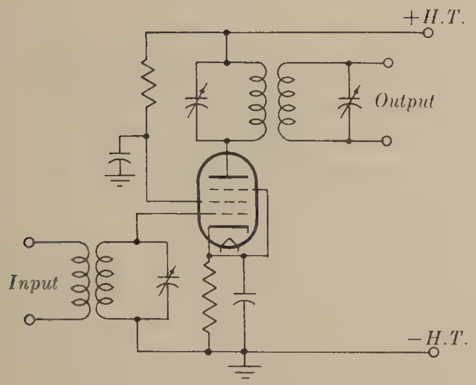


FIG.2-55 Double-tuned transformer coupling.

intermediate frequency is fixed, the i-f amplifiers are tuned so that this frequency lies in the centre of the amplifier band-pass. The adjustment of the i-f transformers to resonate at the fixed i-f is part of the process of aligning the receiver.

The i-f transformer unit consists of the transformer coils and their tuning capacitors mounted in a metal shield. The coils generally used are of two types, air-core and powdered-iron core and are universal-wound. With the air-core transformer, the tuning capacitors are variable and are of the midget air-dielectric type which are more stable in capacitance than other types of variable capacitors. With iron-core transformers, the capacitors may be variable but frequently these are fixed and the inductance of the transformer coils is varied by permeability tuning. Consequently, high-stability fixed mica capacitors are used with this latter type. The stability is important since a circuit with its frequency changing over a period of time has a lower gain because it is not always accurately tuned to the fixed intermediate frequency.

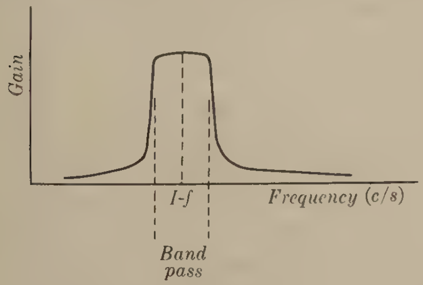


FIG.2-56 Frequency response of a band-pass coupled amplifier.

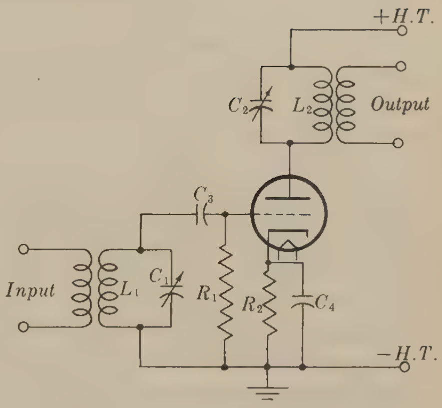
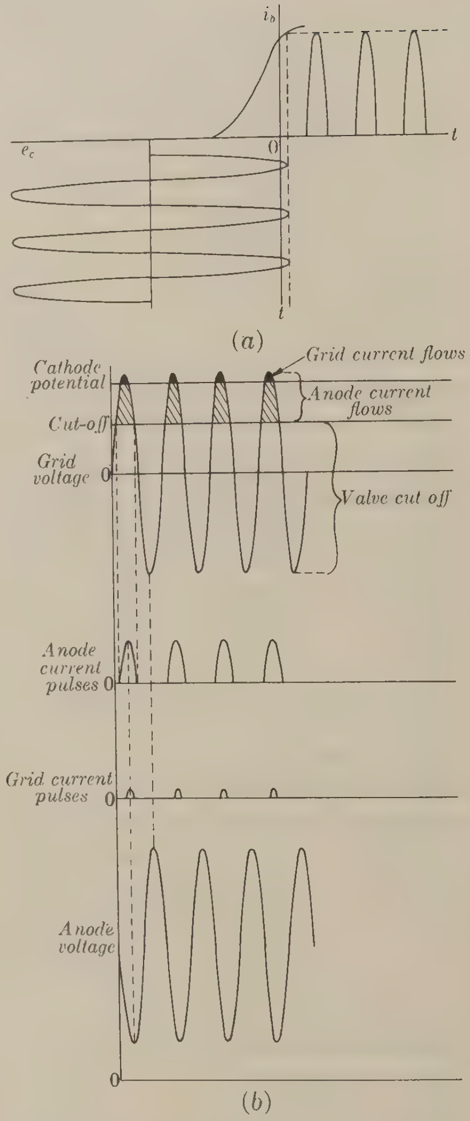


FIG.2-57 Radio-frequency power amplifier.

2.14 Radio-frequency power amplifiers. Radio-frequency power amplifiers find their widest application in transmitters. Triodes, tetrodes or pentodes may be used depending upon the power to be developed. Single-ended circuits, parallel circuits and push-pull circuits may be used as r-f power amplifiers. The circuit diagram of a single-triode amplifier (Fig. 2-57) is very similar to that of the tuned r-f voltage amplifier. The main differences between the two types of amplifiers lie in the physical sizes of the valves, resistors, inductors and capacitors. As the valves operate with a large power output, much heat is developed. They must therefore be of large dimensions and have suitable cooling arrangements.

While Class A or Class B operation is possible, the anode efficiency is so low that this type of operation is ordinarily not employed. Instead, *Class C* operation, which is described below, is almost universally used in r-f power amplifiers. The grid is biased negatively to the cathode by an amount equal to about twice the cut-off voltage. Fixed bias from a separate supply may be used or automatic bias developed in the grid circuit. While cathode bias cannot be used to bias a valve to cut-off, this type is sometimes used in conjunction with automatic bias as shown. The cathode bias provided by C_4R_2 serves to limit the anode current to a safe value when the grid signal is removed and is, therefore, *protective bias.* Additional bias is then provided by C_3R_1 when the signal is applied. The grid circuit L_1C_1 is tuned to resonate at the frequency of the r-f signal which is applied between grid and cathode of the valve. The signal causes the grid potential to vary above and below the bias level as indicated on the graph of the dynamic mutual characteristic in Fig. 2-58 (a). Current flows through the valve only during that part of the cycle when the grid potential is above cut-off. The waveform of anode current is therefore a series of pulses, each of duration less than half of the cycle. These pulses of current flow through the anode circuit L_2C_2 which is tuned to resonate at the signal frequency. The anode current pulses set up an oscillating current in the tuned anode circuit. The waveform of this current and hence of the output voltage which it develops is almost sinusoidal because of the oscillatory effect of C_2L_2. This circuit must not only have a high Q but must present to the valve an impedance equal to the anode resistance of the valve for maximum energy transfer.



(a)

(b)

FIG.2-58 Class C operation of r-f amplifier.

A more complete picture of the operation of the class C amplifier is provided by the waveforms of Fig. 2-58 (b). As previously explained, the anode current pulses flow when the signal drives the grid potential above cut-off. Grid current flows for a short period at the peak of the cycle when the grid is driven positive to the cathode. It is this flow of grid current each cycle which provides the automatic bias. If the grid signal is reduced, less grid current flows and so the bias is reduced. In this way the amplifier sets its own bias level which is proportional to the amplitude of the signal.

The anode voltage is observed to be 180° out of phase with the grid voltage. That is, when the grid becomes most positive, the anode current is a maximum. Hence the p.d. across the tuned anode circuit is a maximum and the anode potential is a minimum. The fact that the grid permits current to flow through the valve only when the anode potential is low makes the anode dissipation relatively low and the anode efficiency high.

Since considerable grid current flows in a Class C amplifier, a greater amount of power must be supplied to the grid circuit than in a Class A amplifier. For this reason the power amplification of the Class C amplifier is relatively low. In large transmitters, therefore, several driver stages are generally employed before the final power amplifier. Each driver in the series operates at a higher power level than the one which precedes it.

2.15 Direct-coupled and cathode-loaded amplifiers. It is noted in section 2.7 that the coupling capacitor in a resistance-capacitance coupled amplifier limits the low-frequency response of the stage. For this reason when d-c voltages or low-frequency a-c voltages are to be amplified it is desirable to dispense with the capacitor and use direct coupling.

Since this means that the grid of one valve is coupled directly to the anode of the preceding one, some readjustment of the d-c levels of the stages is necessary. One of the methods of accomplishing this is to use a battery as coupling between the stages as shown in Fig. 2-59. A battery E protects the grid of V_2 from the d-c component of anode voltage of V_1 and yet allows the grid to follow instantly any change in the anode voltage. Since there are no reactive components in the circuit the voltage gain is independent of frequency at low frequencies. This means that d-c voltages, that is, voltages of zero frequency, are amplified as well as a-c voltages in a direct-coupled amplifier.

Suppose a positive d-c input signal is applied. The anode current of V_1 increases and the voltage drop across R_1 also increases. This drops the voltage at the grid of V_2 and hence the anode current of V_2. Thus the potential at the anode of V_2 rises and maintains its new value as long as the d-c input voltage is applied to V_1. The output voltage is therefore an amplified copy of the d-c input. The disadvantage of this type of direct-coupled amplifier is that the battery voltage changes with age. This not only changes the gain and operating point of the amplifier but may change the entire characteristics of the circuit.

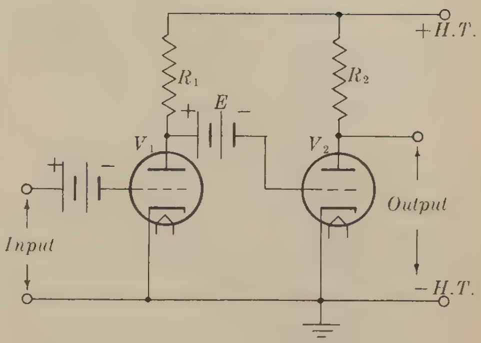


FIG.2-59 Simple type of direct-coupled amplifier.

A direct-coupled amplifier is often used in the measurement of small direct currents and in d-c operated control circuits, relays and telegraph equipments. The direct-coupled amplifier has the following three main disadvantages.

(i) The relatively high-voltage grid-bias batteries are expensive.

(ii) Valve characteristics, battery voltages, or a-c line voltages are subject to changes with time and small changes in the first stage are greatly amplified in the following stages.

(iii) Interelectrode and stray wiring capacitances and the capacitance between grid-bias batteries and the cathodes greatly affect the speed of response to rapid changes of input voltage.

Another type of direct-coupled amplifier in which the whole d-c level of one valve is shifted relative to the other is shown in Fig. 2-60. Here the anode resistor of V_1 is tapped into a voltage

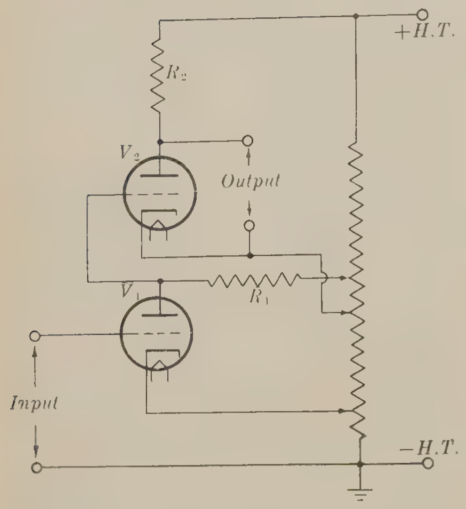


Fig.2-60 A direct-coupled amplifier.

divider just above the cathode voltage of V_2. The voltage drop across the resistor R_1 makes the anode voltage of V_1 and hence the grid voltage of V_2 lower than the cathode voltage of V_2.

Suppose a positive voltage is applied to the grid of V_1. This increases the anode current, drops the anode voltage of V_1 and hence also the grid voltage of V_2. The anode current of V_2 drops, its anode voltage rises and remains high as long as the input voltage is applied to V_1. The voltage rise at the anode of V_2 is the product of the input voltage and the gain of the two stages.

Another type of amplifier which employs direct coupling is the cathode-loaded amplifier. It is an ordinary amplifier with the load resistance placed between cathode and ground as in Fig. 2-61. The input voltage is applied between *grid and cathode* and the output is taken from the cathode load and is in phase with the input.

The valve is biased so that the grid remains negative with respect to the cathode and the valve operates on the linear portion of the mutual characteristic. If a signal is applied and the grid rises in potential the anode current increases. This increase in turn raises the potential of the upper end of R_1. Thus the output voltage is in phase with the input signal voltage. The gain of a cathode-loaded amplifier is the same as that of an ordinary anode-loaded amplifier.

2.16 The cathode follower. In the cathode follower circuit (Fig. 2-62) the load resistor is also placed between the cathode and ground. Unlike the cathode-loaded amplifier however, the input is applied between *grid and ground.* The output voltage is then taken from the cathode load R_k.

In the absence of a signal the current through R_k produces just the bias voltage necessary to permit anode current to flow. If a signal is applied and the grid rises in potential then the current through the valve increases. This increase in turn increases the output voltage across R_k. However this also increases the negative bias. The increase in voltage across R_k is never quite as much as the increase in grid potential so that the amplification is slightly less than one.

Since the change in output voltage tends to counteract the change in input voltage the change in potential of the grid relative to the cathode is the difference between the change in grid voltage and the change in voltage across R_k. This is very much smaller than the change

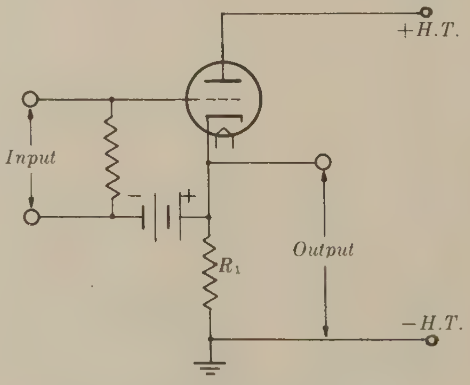


Fig.2-61 Cathode-loaded amplifier.

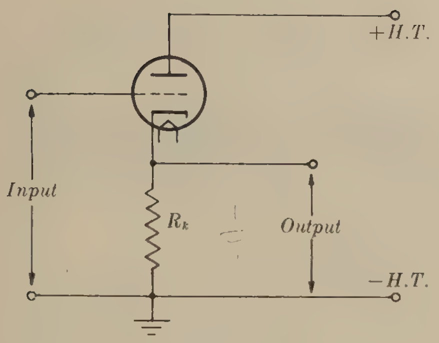


FIG.2-62 Cathode follower.

in grid-signal voltage and consequently the cathode follower can handle very large grid-signal amplitudes without driving the grid positive and introducing distortion.

The name cathode follower is derived from the fact that the output voltage is in phase with the input voltage. The cathode follower is an inverse feedback amplifier since the output voltage appears in the input in opposite phase to the grid-signal voltage.

An expression for the voltage gain of a cathode follower in terms of the valve and circuit constants is

$$A = \frac{\mu R_k}{r_a + R_k(\mu+1)}$$

or

$$A = \frac{1}{\dfrac{r_a+R_k}{\mu R_k}+1}$$

This shows that A is always less than 1 since $\dfrac{r_a+R_k}{\mu R_k}$ is never zero, but A is always close to 1.

Even though there is no amplification in a cathode follower it has certain characteristics which make it very useful.

The input impedance of the cathode follower is very large since the grid is never driven positive. The effect of interelectrode capacitances in the valve is also greatly reduced. For a cathode follower the input capacitance is $C_{ga}+C_{gk}(1-A)$ but in a cathode follower A is at most 1, and therefore C_{gk} has hardly any effect. Thus the input capacitance is reduced approximately to C_{ga}. This very high input impedance is useful since it does not load the input circuit.

In an ordinary amplifier the band of frequencies passed is determined by the input impedance of the amplifier stage. Since this is a capacitive impedance it decreases greatly as the frequency is increased. As pointed out above, the input impedance of the cathode follower is very high, since the input capacitance is very low. Thus the signal frequency can be very high without the impedance changing to any great extent. The cathode follower will pass a band of frequencies from 10 c/s or lower to 4 Mc/s or higher. This makes it very useful in passing on without distortion pulse voltages, such as square waves, which contain a great number of frequencies.

The output impedance of the cathode follower is very small or, in other words, the internal resistance of the output circuit is very small. The internal resistance of any source of power shows itself as a drop in the p.d. across the output terminals as more current is drawn. An increase in load current would cause a drop in p.d. across the terminals of R_k. However, this lowering of voltage across R_k permits more current to flow through the valve. This increased current causes a rise in p.d. across R_k to very nearly its original value. Thus the internal resistance of the cathode follower is small and it can therefore feed maximum power to the low impedance of a coaxial cable, a speaker or any low-impedance load without any additional matching device.

The cathode follower has the following characteristics:

(i) an amplification which approaches but never reaches 1

(ii) a very high input impedance over a wide frequency band

(iii) a very low output impedance over a wide frequency band

(iv) a uniform frequency response and low phase shift over a wide frequency band

(v) an output which is in phase with the input.

Ordinarily the cathode follower is operated without an external source of grid bias. When this is true, as the signal is increased in amplitude, eventually the negative peaks are clipped because the valve is driven to cut-off. There is still no danger of this large signal amplitude driving the grid positive because, as the grid goes more positive, more current flows to cause a larger voltage drop across R_k and thus a larger bias. Therefore in order to prevent clipping an additional steady positive bias may be applied to the stage. This allows a larger amplitude signal to be handled without distortion.

2.17 Video-frequency amplifiers. Wide band-pass amplifiers which are capable of handling complex waveforms without distortion are called *video amplifiers* because they are generally used to amplify signals from which visual intelligence can be derived by means of a cathode-ray tube. Video amplifiers find application in oscilloscopes, in ionosphere sounding, in television, and in various pulse circuits.

A. General Discussion

A voltage with a complex waveform is the sum of a fundamental sine-wave voltage of the same period and a large number of harmonics. Examples of such voltages are the saw-tooth voltage used in a linear time base and a square-wave voltage which is used in certain pulsing circuits. For example, if the fundamental or first harmonic (1°) is combined with those of the odd harmonics, 3°, 5°, 7°, 9°, as shown in Fig. 2-63 (a), the resultant waveform obtained has a fairly square shape. The repetition rate of the resultant wave is equal to the frequency of the fundamental. Further increase in the odd harmonic content produces a waveform which is more nearly square as shown in Fig. 2-63 (b).

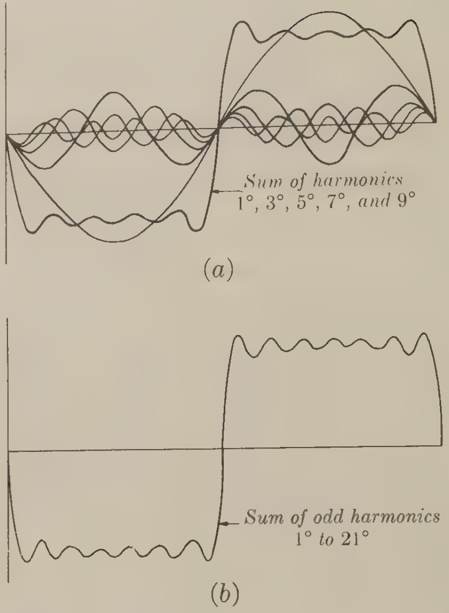


FIG. 2-63 Square waveform (a) Fundamental and 5 harmonics (b) Fundamental and many harmonics.

Fig. 2-64 (a) and (b) shows how a saw-tooth wave of the type useful for linear sweep circuits can be approximated by combining eight and then twenty harmonics. The ideal form of both the square wave and the saw-tooth wave consists of an infinite number of harmonics of the fundamental frequency.

If these waveforms are to be passed by an amplifier without distortion, the amplifier must have an exceptionally wide band-pass and an angular phase shift proportional to the frequency. The ideal frequency-response curve is that shown in Fig. 2-65. Phase distortion which can be neglected in audio work because the ear does not distinguish between moderate changes in phase, becomes an important consideration in video amplifiers since

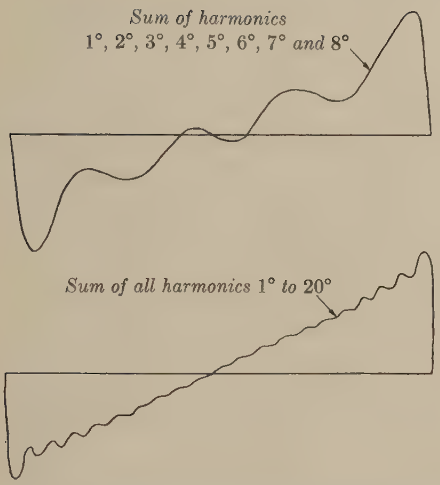


Fig.2-64 Sawtooth waveform (a) Fundamental and 8 harmonics (b) Fundamental and 20 harmonics.

the cathode-ray oscilloscope is sensitive to minute phase shifts. The ideal phase characteristic is shown in Fig. 2-66 where the phase shift angle is proportional to the frequency.

In practice, small changes in wave shape brought about by failure to pass harmonics of extremely high order do not seriously affect the quality of the output. In fact from the standpoint of signal-to-noise ratio there is some advantage in restricting the bandwidth. Optimum signal-to-noise ratio is obtained when the band-pass is twice the frequency whose period is equal to the pulse width. Thus for one-microsecond pulses a two-megacycle bandwidth gives optimum signal-to-noise ratio. However, if the pulse repetition rate is 500 c/s all harmonics above the 2000th are suppressed with this bandwidth. This means that optimum signal-to-noise ratio is achieved at the expense of high-frequency response. It should be noted that the higher the order of harmonics passed the more perpendicular are the edges of the output wave for a square-wave input. The deviation from the perpendicular permitted by a certain application sets the upper limit of frequency which must be passed. The repetition rate sets the low-frequency limit.

The three principal requirements of a video amplifier which distinguish it from an audio amplifier are:

(i) a constant frequency response over a band several Mc/s wide
(ii) a constant time delay or phase shift proportional to frequency over the usable range
(iii) a satisfactory signal-to-noise ratio while attaining the first two requirements.

Because of the very wide frequency response requirements of the video amplifier, only resistance-capacitance coupling and cathode coupling are suitable. If high gain is to be achieved cathode-follower stages can not be used because their stage gain is always less than unity. The cathode follower does have its place in video amplifiers, however, as a high-impedance input stage, as a buffer between high-gain stages and as a matching device when a video ampli-

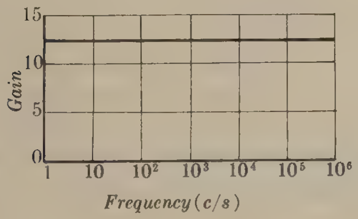


Fig.2-65 Frequency response of an ideal video amplifier.

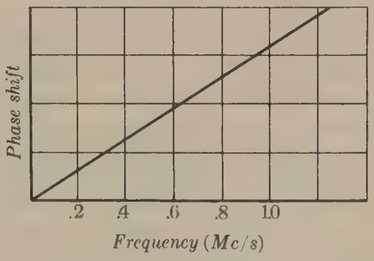


Fig.2-66 Phase shift of an ideal video amplifier.

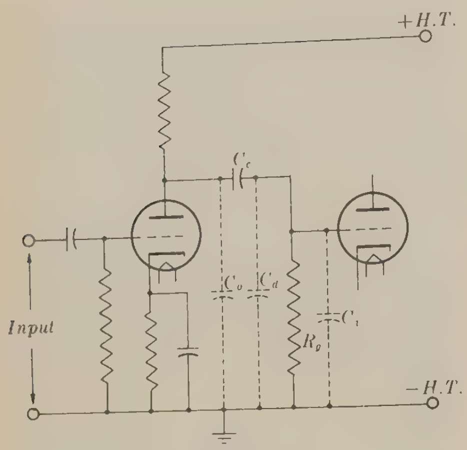


FIG.2-67 Video amplifier circuit.

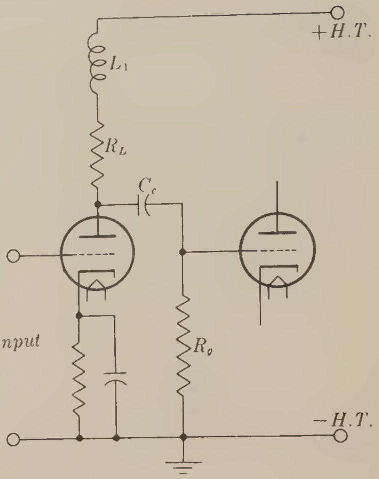


FIG.2-68 High-frequency compensation, shunt type.

fier is coupled to a low-impedance load such as a transmission line.

A video amplifier using a resistance-capacitance coupled circuit differs from an audio amplifier only in the values of the circuit components. Several difficulties are encountered when the amplifier shown in Fig. 2-67 is used as a video amplifier. The high-frequency response is limited by the reduction of the load impedance caused by the shunting effect of the output capacitance C_o, distributed capacitance C_d, and input capacitance C_i of the following stage. The low-frequency response is limited by the time constant R_gC_c which must be long compared with the period of the lowest-frequency voltage to be amplified. Valves with very low interelectrode capacitance must be used in video amplifiers. The circuit must be carefully wired with short, properly spaced leads to reduce distributed capacitance in the wiring.

The term *video amplifier* is sometimes applied to high-gain tuned amplifiers with a band-pass as narrow as 80 Kc/s. The application of the term is justified when an amplifier is used to amplify a signal which is rendered intelligent by application to the deflection plates of a c.r.t.

B. High-Frequency and Low-Frequency Compensation

It is possible to compensate for the shunting effect of the interelectrode and distributed capacitances and to raise the upper frequency limit of the amplifier by connecting a small inductance L_1 in series with the load resistor R_L (Fig. 2-68). This *shunt-compensation coil* increases the high-frequency gain by resonating with the shunt capacitances at a frequency above the normal upper frequency limit. That is, the load impedance is made to increase by the resonating effect to counteract the shunting effect of the capacitances which increases at high frequencies. Thus, the flat part of the response curve is extended up to the resonant frequency of L_1 and the shunt capacitances, without decreasing the mid-frequency gain, as shown in Fig. 2-71. This shunt compensation method is effective when the upper frequency limit is not too high and there are only a few stages of amplification.

Another method of high-frequency compensation is *series compensation*. Fig. 2-69 shows a small compensating inductance L_2 connected in series with the coupling capacitor. This series compensa-

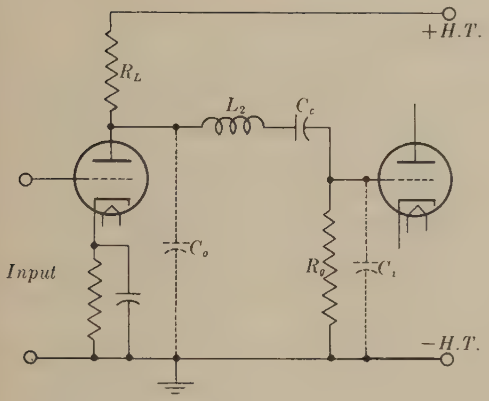


FIG.2-69 High-frequency compensation, series type.

tion coil increases the high-frequency gain by resonating with C_i and hence producing an increased voltage across it.

These two methods can be combined in a shunt-series compensation circuit (Fig. 2-70) which has the high-frequency peaking effect of the first method as well as the increased gain of the second.

Sometimes the upper frequency limit is increased merely by reducing the size of R_L. This has the disadvantage of reducing the gain but at the same time the effective input capacitance, $C_{gk}+C_{ga}(1+A)$ is decreased. The reactances of the shunt capacitances are then larger relative to the load resistance and the shunting effect across R_g is reduced.

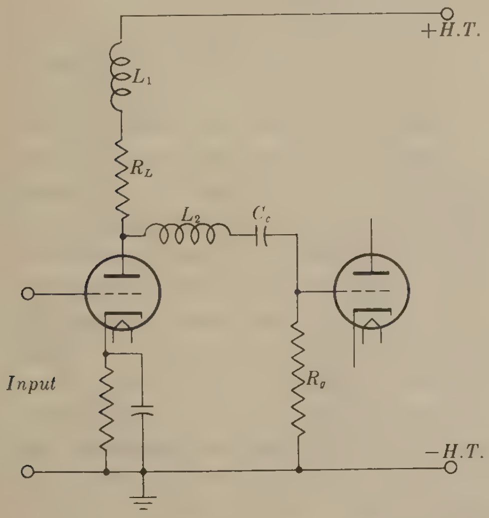


FIG.2-70 High-frequency compensation, series-shunt type.

At low frequencies the gain of a video amplifier is reduced by the coupling capacitor. This reduction in gain is due to the increasing reactance of the capacitor at frequencies below about 200 c/s. The value of C_c is chosen large enough for the reactance at the middle and high frequencies to be negligible and the full output voltage appears across R_g. As the frequency decreases the reactance of C_c increases and less of the output voltage appears across R_g. This results in less of the output of one stage appearing at the input of the following stage. This effect is satisfactorily reduced in some applications by the choice of a large coupling capacitor and a large grid leak in the following stage so that most of the output voltage, even at low frequencies, appears across the grid resistor. The extent to which the value of the coupling capacitor can be increased is limited since an increase in the physical size of the capacitor results in an increase in the stray capacitance which affects the high-frequency response. For this reason the most effective additional means of compensation is necessary.

In cases where the low-frequency gain is not sufficient, a *low-frequency compensation filter*, consisting of a resistor R_f and a capacitor C_f in parallel, is connected in series with the load resistor (Fig. 2-72).

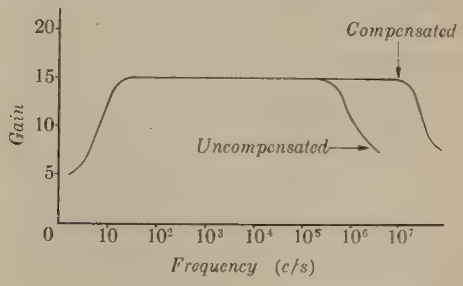


FIG.2-71 Effect of high-frequency compensation on response curve.

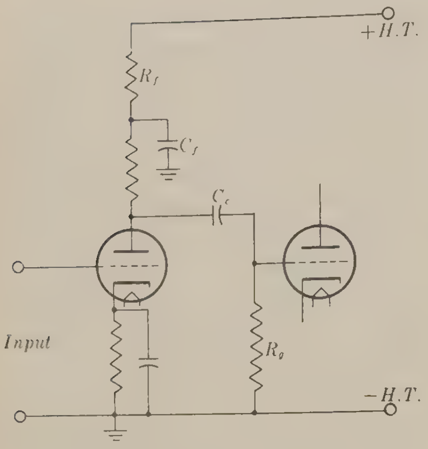


FIG.2-72 Low-frequency compensation.

This filter introduces a phase lag to offset the phase lead introduced by the coupling capacitor. It also effectively increases the anode load impedance at low frequencies which maintains the gain and extends the low frequency limit. This effective increase of load impedance is accomplished by an increase in the reactance of C_f at low frequencies. An alternative method of extending the low-frequency response down to zero frequency or d-c is to use direct-coupled amplifiers.

C. Valves Suitable for Video Amplifiers

The degree to which the high-frequency response of a video amplifier is extended is dependent to a large extent on the characteristics of the valves used. The most important factors in the choice of valves are the *transconductance* and the *interelectrode capacitances*, mainly C_{ga}. An approximate *figure of merit* for a valve is obtained by dividing the transconductance by the value of C_{ga} and comparisons of valves are made on the basis of this ratio. This comparison is of value provided no variations of total capacitance are introduced by stray wiring capacitance when the valves are used in practical circuits.

For a triode the total input capacitance is given by

$$C_i = C_{gk}+C_{ga}(1+A)$$

If the gain of the stage is high, the input capacitance is large owing to the factor $A\times C_{ga}$. For this reason triodes are not used in video amplifiers except in the input stage where they are useful because of their low inherent noise level. A typical value of the figure of merit for a triode is 50.

On account of its lower interelectrode capacitances a pentode is a more satisfactory valve in spite of the fact that it may have a lower value of g_m. For example, a typical value of the figure of merit for a pentode is 100.

Special television pentodes have been developed for video amplifiers. These have a very high value of g_m which is attained by using a very small spacing between the control grid and cathode and a very fine winding pitch on the spirals of the grid. Unfortunately, increasing g_m by this method results in an increase in the interelectrode capacitance of the valve. However, the increase in g_m is several times the increase in interelectrode capacitance and the figure of merit of a television pentode is increased to between 350 and 550.

D. Video Power Amplifiers

When a video amplifier must supply a high voltage output to a circuit with a low input impedance, considerable power is drawn. Ordinary voltage amplifier pentodes are not capable of supplying sufficient power output without excessive harmonic distortion. Power pentodes or beam power valves are used instead. An example of a video power stage is the modulator stage of a transmitter where the modulation is a video signal.

When the load on a video amplifier is a low impedance such as a coaxial cable it is important that the output impedance of the final amplifier stage be matched to the load. Matching is accomplished for loads of 500-2000 ohms

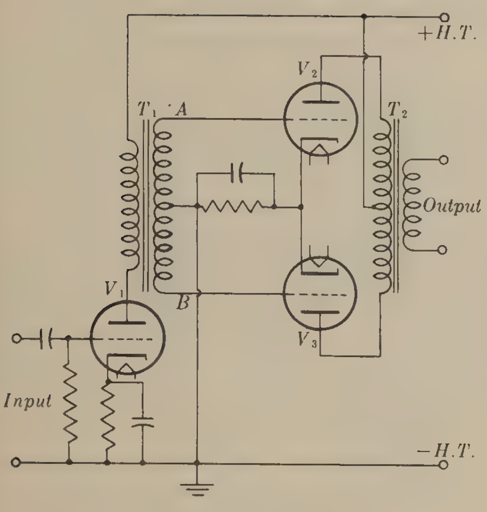


FIG. 2-73 Phase-splitting circuit.

by using a low value of anode load resistor approximately equal to the impedance to be matched. Although the stage gain is lowered the frequency response is not changed.

A more suitable circuit for an output stage in which low impedance is required is the cathode follower. In addition to low output impedance it features good stability and freedom from distortion. The cathode follower stage is especially useful for coupling grounded-sheath coaxial cable circuits because one end of the output impedance is at earth potential and no blocking capacitor is required at the other end. The high input impedance of the cathode follower stage makes possible the use of a large coupling capacitor and a small value of grid resistor which improves the low-frequency response without affecting the high-frequency response. Furthermore, since the cathode follower is an inverse feedback circuit, non-linear distortion is reduced and much of the instability due to fluctuations in the line voltage and changes in valve constants is eliminated.

2.18 Phase-splitting circuits and paraphase amplifiers. The advantages of operating an amplifier stage push-pull rather than single-sided are discussed in section 2.12. In order to couple the a-c output voltage of a single valve to a push-pull circuit there must be developed in the coupling circuit two voltages equal in magnitude and 180° out of phase. A circuit which provides this push-pull output from a single input waveform is called a *phase splitter*.

One of the most common phase-splitting circuits consists of a transformer with a centre-tapped secondary. Fig. 2-73 shows a single-sided amplifier coupled to a push-pull amplifier by means of the transformer T_1. With the centre tap grounded the point B goes positive at the same instant that the point A goes negative with respect to ground and the voltage on the grid of V_2 is of opposite polarity to the voltage on the grid of V_3. If the grounded tap is exactly in the centre of the secondary winding, the amplitudes of the respective grid voltages are equal.

Sometimes it is difficult to balance the secondary exactly. In this case the alternate arrangement of Fig. 2-74 may be used. Since the voltage across the resistors R_1 and R_2 is equal to the voltage across the secondary of the transformer, grounding the junction of the resistors

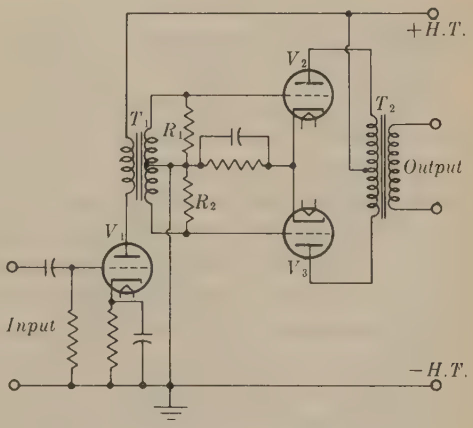


FIG. 2-74 Balanced phase-splitting circuit.

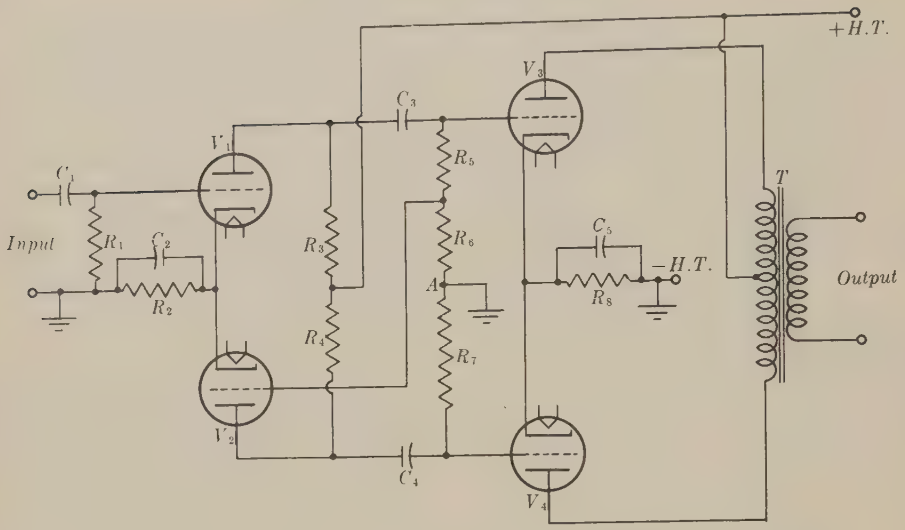


FIG.2-75 Paraphase amplifier.

effectively grounds the mid-point of the transformer winding if the resistors are equal in size. The effect of this circuit is the same as that of a transformer with a centre-tapped secondary winding. In this case, however, some care is required in the choice of resistance for R_1 and R_2, since these two resistors form part of the load impedance on the valve V_1. The reflected load resistance in the anode circuit of V_1 due to the resistors R_1 and R_2 is given by

$$\frac{R_1+R_2}{\frac{N_s}{N_p}}$$

where $\frac{N_s}{N_p}$ = transformer turns ratio.

Thus, if T_1 is a step-up transformer the reflected load resistance is much smaller than the actual series resistance of R_1 and R_2 determined by the permissible grid resistance for V_2 and V_3. Since the grid circuits of this push-pull amplifier contain more resistance than those using a centre-tapped transformer, this type of amplifier gives more distortion if the grids are driven positive.

Transformer coupling is expensive, attenuates low frequencies and gives high-frequency distortion due to the variation in transformer ratio with frequency. In addition the losses of the transformer vary with frequency.

These undesirable features are avoided if the transformer coupling is replaced by resistance-capacitance coupling. The input voltage is applied to an amplifier to obtain the desired amplitude of voltage input for one of the push-pull valves. Some of the output of the amplifier is then applied to a second amplifier to provide a voltage of the same amplitude but of opposite polarity to the output of the first amplifier circuit. The output of this second amplifier is used as an input for the other push-pull valve. The second amplifier is called a *phase inverter*. This combination of amplifier and phase inverter used to provide a push-pull output from a single input voltage is called a *paraphase amplifier*, or simply a *phase inverter*.

A two-valve paraphase amplifier consisting of amplifier and phase inverter is shown in Fig. 2-75. To obtain the desired amplitude of output, V_1 is connected as a conventional amplifier. The output voltage of V_1 appears across the

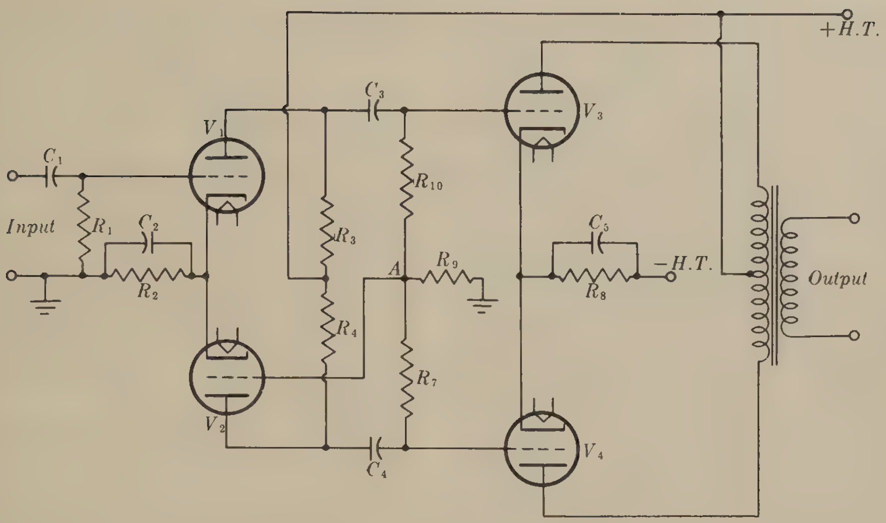


FIG.2-76 Floating paraphase amplifier.

voltage divider R_5 and R_6 and is fed directly to the grid of V_3. The portion of the output of V_1 appearing across R_6 is amplified by the phase inverter V_2 to the same amplitude as that of the output of V_1 and fed to the grid of V_4.

When an increased signal swing in the positive direction causes increased current through V_1 the voltage drop across the anode load resistor R_3 increases and the anode potential drops. This drop in potential is coupled through C_3 to the grid of V_3 across the voltage divider R_5, R_6. Thus the grid of V_3 is made more negative. At the same time the portion of voltage drop across R_6 is applied to the grid of V_2 where it causes a decrease in anode current and the anode of V_2 therefore goes more positive. This increase in potential is coupled via C_4 to the grid of V_4. Thus the grid of V_4 swings in the positive direction at the same time as the grid of V_3 swings in the negative direction and, in the middle range frequencies, the desired 180° phase difference for the push-pull grids is achieved. At frequencies above and below this range where the capacitive reactances have an appreciable effect the two output voltages are not exactly 180° out of phase.

The circuit is adjusted so that equal voltages are applied to each grid of the push-pull stage. In the case where identical valves or a twin valve are used for V_1 and V_2, R_3 should equal R_4 and R_7 should equal the sum of R_5 and R_6. The size of R_6 should be the same fraction of the sum of R_5 and R_6 as the input voltage to V_1 is of the output voltage of V_1.

That is, $$R_6 = \frac{R_5+R_6}{A}$$

where A = the stage gain of V_1.

Thus the voltage input to V_2 is reduced to that value, which after amplification by V_2 gives a voltage equal to the output of V_1.

A self-balancing type of paraphase amplifier known as the *floating paraphase* amplifier employs the differential voltage between the outputs of the two valves as the input signal to the phase-inverter section (Fig. 2-76). The circuit of Fig. 2-75 may be modified to produce this type of paraphase amplifier by connecting between the point A and ground a resistor R_9 which has the same value as each of R_7 and R_{10}. The valve V_1 is an amplifier used to obtain the required amplitude of input voltage to the grids of the push-pull circuit. The output voltage of V_1 is applied across the resistors R_{10} and R_9. The output voltage of V_2 is applied across R_7 and R_9. Therefore half the output of both V_1 and V_2 is applied across R_9. Since these two output voltages are always of opposite polarity, the resultant voltage across R_9 is half the difference between the outputs of V_1 and V_2. This is the voltage which is applied to the grid of V_2. In order to obtain the proper phase relations between the output voltages of this paraphase amplifier, the output voltage of V_1 must be slightly larger than that of V_2. This is because the voltage at the grid of V_2, which is half the difference between the outputs of the two valves, must be in phase with the output of V_1. Since this difference should be kept as small as possible, pentodes are usually used on account of their large amplification factor. In this type of phase-inverter feed, the voltage gain for valves V_1 and V_2 may vary considerably without introducing much unbalance between the grid voltages fed to the push-pull stage. This reduces the effect of differences between valves V_1 and V_2 that may arise as the valves age through use.

Although the input voltage to V_3 is always slightly greater than the input voltage to V_4 this difference is not sufficient to interfere with the proper operation of the push-pull amplifier. Any greater difference is counteracted as follows. The increased difference voltage across R_9 is amplified by V_2 to provide a greater input voltage to V_4. This reduces the difference between the voltages applied to the grids of V_3 and V_4.

2.19 Inverse-feedback amplifiers. If some of the energy from the anode circuit of an amplifier is coupled back into the grid circuit, the amplifier has *feedback*. If the voltage fed back from the anode circuit is such that it aids the applied signal, the feedback is *positive* or *regenerative*. That is, a positive feedback voltage has a component which is in phase with the applied signal. If the voltage fed back is such that it opposes the applied signal, the feedback is *negative, inverse* or *degenerative*. That is, a negative feedback voltage has a component which is 180° out of phase with the applied signal.

If positive feedback or regeneration exists in a circuit, the total amplification is increased. This is because the amplitude of the signal at the grid has been increased by the feedback and this larger signal is amplified in the same ratio, giving a greater output voltage than if the signal voltage alone were applied to the grid. Unfortunately, increasing the gain of the amplifier amplifies and exaggerates any undesirable distortion or noise which may be introduced by the valve itself. Furthermore, a circuit with positive feedback tends to amplify selectively at one frequency, since the feedback voltage tends to become a maximum at the frequency at which the original amplification is greatest. It therefore increases the selectivity of the amplifier, and hence is used chiefly where high gain and sharpness of resonance both are wanted. Sometimes the regeneration becomes so large that sufficient energy is fed back to maintain the operation of the circuit without the usual input voltage and the amplifier becomes an oscillator. This case is discussed in Chapter V on oscillators.

In a negative-feedback or degenerative circuit the amplification is decreased, since the grid-signal voltage is reduced and hence the amplified output voltage is smaller than if the signal voltage alone were applied to the grid. The effect of negative feedback is to make the fre-

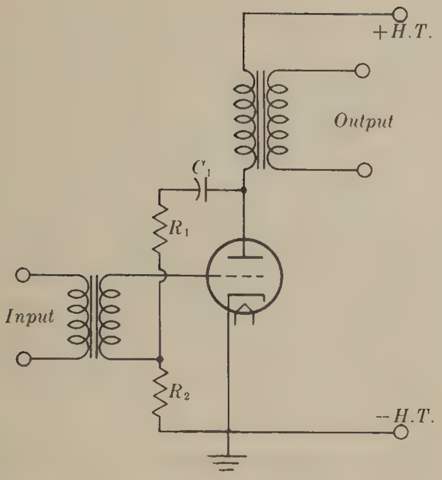


FIG.2-77 Negative feedback from anode to grid circuit.

quency response more uniform by reducing the gain more in the middle-frequency range than in the upper or lower frequency range where the gain is lower to begin with. Negative feedback also improves the waveform at the output of an amplifier by reducing the harmonic distortion which arises within the amplifier. The signal which is applied to the grid is amplified by a factor A in the output, but the harmonic distortion caused by the non-linear characteristic of the amplifier is generated within the valve and is therefore not amplified in the output. Consequently when a fraction of the output is fed back out of phase with the input, the distortion component of the feedback voltage is amplified by the same factor as the input signal. As a result, the net output may contain a minimum of distortion, but the amplitude of the desired signal is considerably reduced. However, negative feedback has no effect on the harmonic distortion caused by the flow of grid current in the input stage of an amplifier since this occurs at the grid, and the distortion is therefore amplified in the same ratio as the desired signal.

In the same manner, negative feedback reduces noise arising within the amplifier. However, if the noise is present in the signal it will be amplified as well as the signal itself. Both positive and negative feedback may be applied over several stages of an amplifier, rather than from the anode circuit to the grid circuit of a single stage.

Negative feedback may be obtained in an amplifier in several ways. One method is to couple a portion of the output voltage into the grid circuit by means of a voltage divider (Fig. 2-77). When the grid potential is made less negative, the anode current increases and the lower end of the primary of the output transformer drops in potential. This drop in potential is transferred through the capacitor C_1 to the voltage divider R_1 and R_2 and appears at the upper end of R_2, thus adding an increasing negative feedback voltage to the increasing positive grid-signal voltage through the secondary of the input transformer.

Another method of providing negative feedback is to use a cathode resistor which is not by-passed (Fig. 2-78). In this case the flow of anode current through the cathode resistor R_k develops a feedback voltage across it. This voltage appears in series with the grid-signal voltage on the grid of the valve since the cathode resistor is placed between the

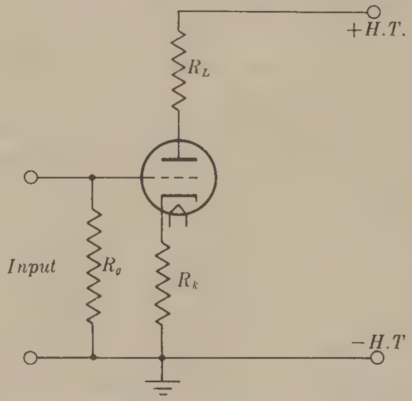


FIG.2-78 Negative feedback in the cathode circuit.

cathode and ground. When the grid potential is made less negative the resulting increase in anode current causes a rise in potential at the upper end of R_k which is connected to the cathode. This is equivalent to a drop in the potential of the grid relative to the cathode, opposing the grid-signal voltage. This is called *current feedback* as contrasted with voltage feedback shown in Fig. 2-77. The cathode follower circuit and some types of phase inverters employ this type of inverse feedback.

Inverse voltage feedback is the type in which the feedback voltage is proportional to the output voltage appearing across the load. Voltage feedback reduces the effective internal resistance of the amplifier since the feedback voltage is usually obtained from a voltage divider in parallel with the anode load. Therefore this type of feedback is useful when a large amount of power must be supplied to the load.

Negative feedback may be applied over more than one stage if the proper phase relations are developed. Fig. 2-79 shows a two-stage amplifier with inverse feedback. The feedback voltage is developed across the resistor R_2 which, in series with R_1, forms a voltage divider for feeding back the desired fraction of output voltage of the second stage to the grid of the first stage. If the grid of the first valve goes more negative the grid of the second goes less negative. The resulting increase in anode current causes the potential at the anode of V_2 to drop. This drop in voltage is transmitted through R_1 and C_1 to the upper end of R_2, causing a drop in potential at this point. Thus, the potential of the cathode of V_1 is made less positive with respect to ground, thus making the grid of V_1 less negative to the cathode. Hence the feedback voltage decreases the amplitude of the negative signal voltage.

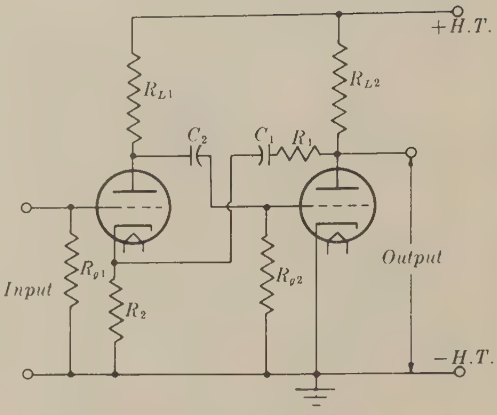


FIG. 2-79 Negative feedback in a two-stage amplifier.

The gain A' of an amplifier with feedback may be found in terms of the gain A without feedback and the *feedback factor* β, where β is the fraction of the output voltage which is fed back to the input. Positive feedback is indicated by giving β a positive sign and negative feedback by a negative sign.

If the a-c output voltage is e_a, a portion βe_a of this output is fed back out of phase with the input signal voltage e_g. Therefore the total input voltage $e_i = e_g - \beta e_a$. The gain of the amplifier without feedback is given by

$$A = \frac{e_a}{e_g} \qquad (1)$$

The gain of the amplifier with feedback is given by

$$A' = \frac{e_a}{e_i}$$

$$\therefore\ A' = \frac{e_a}{e_g - \beta e_a}$$

But from (1) $e_a = Ae_g$

Therefore $$A' = \frac{Ae_g}{e_g - \beta A e_g}$$

or $$A' = \frac{A}{1 - \beta A}$$

where A' = gain with feedback
A = gain without feedback
β = feedback factor.

If β is positive the quantity $(1 - \beta A)$ is less than 1 and the gain of the amplifier is increased by positive feedback. If the quantity βA is made equal to 1 the denominator of the expression is zero and the positive feedback is large enough to maintain the operation of the system unaided by a signal input. This is the principle of the oscillator circuit.

If β is negative the quantity $(1-\beta A)$ is greater than 1 and the gain of the amplifier is reduced. If βA is made much greater than 1 the quantity $(1-\beta A)$ is very nearly equal to $-\beta A$. Therefore the gain of the amplifier with negative feedback is $\frac{1}{\beta}$ in value since β has negative sign. If the feedback is obtained through a resistive network, the feedback factor β is independent of frequency and the gain is uniform over a wide range of frequencies. If a different frequency-response curve is desired the feedback circuit must be designed with the proper components so that the factor β depends on frequency.

If the load impedance of an inverse-feedback amplifier is decreased the output voltage decreases. If the load is separate from the feedback path, the feedback voltage also decreases, causing the effective gain to increase. This increased gain restores the output voltage nearly to its former value. On the other hand, if the load is increased the resulting rise in output voltage is offset by the increased feedback and the gain remains substantially the same. Thus, if the load impedance does not form part of the feedback path, the gain of the amplifier does not change with changing load impedance. It is shown above that the gain of an inverse feedback amplifier is equal in value to $\frac{1}{\beta}$. Thus the gain is not affected by variations in supply voltages or by ageing of the valves. Hence the gain is more stable in an inverse-feedback amplifier if the feedback is large.

2.20 Impedance matching. The amount of power consumed by a load circuit depends upon the voltage applied to the load, the current which flows in the circuit and the phase relationship between the current and the voltage. In order to cause a specified amount of power to be consumed by the load circuit, the impedance of the load must be adjusted to a suitable value.

If, in a particular application, the amount of distortion which the circuit introduces is of no importance, maximum transfer of energy from the source to the load occurs when the impedance of the load is made equal to the internal impedance of the source. The process of adjusting the impedance of the load to the internal impedance of the source is called *impedance matching.*

If the distortion which the circuit introduces must be kept from exceeding a given amount, the impedance of the load circuit may have to be adjusted to a value somewhat different from the internal impedance of the source. In this case, maximum transfer of energy is no longer obtained. The process of adjusting the impedance of the load to effect maximum transfer of energy within the limits of permissible distortion, is called *optimum matching.*

A. Impedance Matching for Maximum Power Output

The significance of impedance matching for maximum power transfer from the source to the load is readily understood by considering the operation of the circuit shown in Fig. 2-80. A source of e.m.f. of 100 volts having an internal resistance of 10,000 ohms is used to

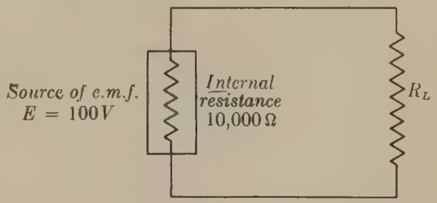


FIG.2-80 Circuit to show effect of load impedance on power output.

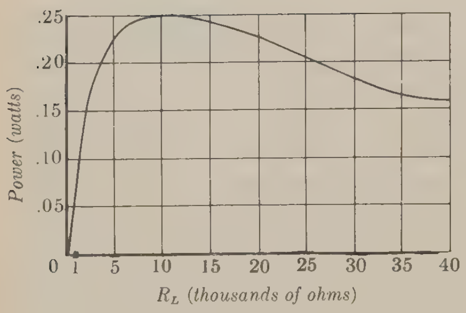


Fig.2-81 Graph of impedance matching.

supply power to the load circuit R_L. Such a circuit may represent a battery supplying power to a resistor, a valve connected to a loudspeaker, a transmitter feeding an aerial or any similar application. Various values of load impedance are used in the circuit and the power consumed by the load is calculated in each case. The power consumed in the load circuit is equal to the product of the current through the load and the voltage across it. The current that flows in the circuit is determined from Ohm's Law as the e.m.f. of the source divided by the total resistance of the source and the load. The voltage across the load is also determined from Ohm's Law as the product of the current flowing in the circuit and the resistance of the load. The values of current, voltage and power in the load circuit are given in the following table.

The variation in power consumed by the load with various values of load resistance is further illustrated by the graph of Fig. 2-81. It should be noted from the table and from the graph that the power consumed in the load is a maximum when the resistance of the load equals the internal resistance of the source which is 10,000 ohms in this example. If the resistance of the load is made less than the resistance of the source, the current that flows in the circuit is increased but the voltage is decreased. That is, the load exerts a short-circuiting effect on the source and the reduced voltage results in the load power being less than the maximum. On the other hand if the load resistance is large in comparison with the internal resistance, the voltage applied to the load is large but the current flowing in the circuit is small. This also results in the external power being less than the maximum power. Maximum power transfer from the source to the load occurs when the internal and external resistances are matched. This principle is true for impedance loads as well as resistance loads.

If the load impedance and the internal impedance are not equal, the circuit is said to be *mis-matched*. It is evident from Fig. 2-81 that the power output of the source falls off very steeply if the load impedance is less than the internal impedance. The reduction in output

LOAD RESISTANCE	TOTAL RESISTANCE	CURRENT	LOAD VOLTAGE	LOAD POWER
1000 ohms	11,000 ohms	$\frac{100}{11,000}A$ $= 9.09mA$	$\frac{100}{11,000} \times 1000$ $= 9.09$ volts	9.09×9.09 $= .083$ watts
5000	15,000	6.67	33.3	.222
10,000	20,000	5.00	50.0	.250
15,000	25,000	4.00	60.0	.240
20,000	30,000	3.33	66.7	.222
30,000	40,000	2.50	75.0	.187
40,000	50,000	2.00	80.0	.160

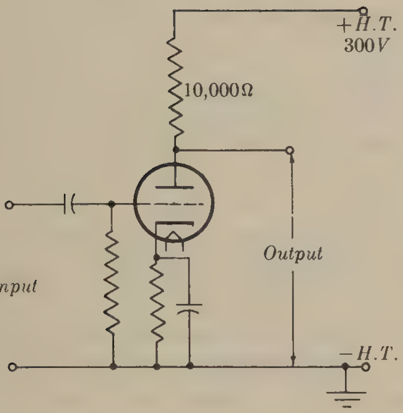


FIG.2-82 Triode amplifier with load equal to anode resistance.

power is less marked when the load impedance is greater than the internal impedance. If a mis-match is unavoidable therefore, the load impedance should be made larger than the internal impedance to maintain a large output.

B. *Impedance Matching for Optimum Power Output*

A graphical analysis of the operation of an amplifier (Fig. 2-82) shows how the power output and distortion vary with different values of load impedance. The valve is a low-mu triode connected in a resistance-capacitance coupled voltage amplifier. The effect of connecting a load resistor equal to the anode resistance of the valve (10,000Ω) is shown in Fig. 2-83. The dynamic e_c-i_b curve and the load line (e_b-i_b curve) are drawn for the amplifier circuit. The grid bias of −8 volts is marked off and the operating point *P* identified. A sinusoidal signal of 6 volts amplitude is superimposed upon the steady grid bias on the e_c-i_b characteristic. The anode current is observed to vary from a minimum of $2mA$ to a maximum of $13mA$. These values of anode current are carried over to the load line and the corresponding values of anode voltage are noted as 280 volts and 170 volts respectively. Thus a total grid voltage swing of 12 volts produces a corresponding anode current swing of $11mA$ and an anode voltage swing of 110 volts.

The peak value of a-c component of anode current,

$$I_{a\ max} = 5.5mA$$

∴ the effective value of the a-c component of anode current,

$$I = \frac{5.5}{\sqrt{2}}mA.$$

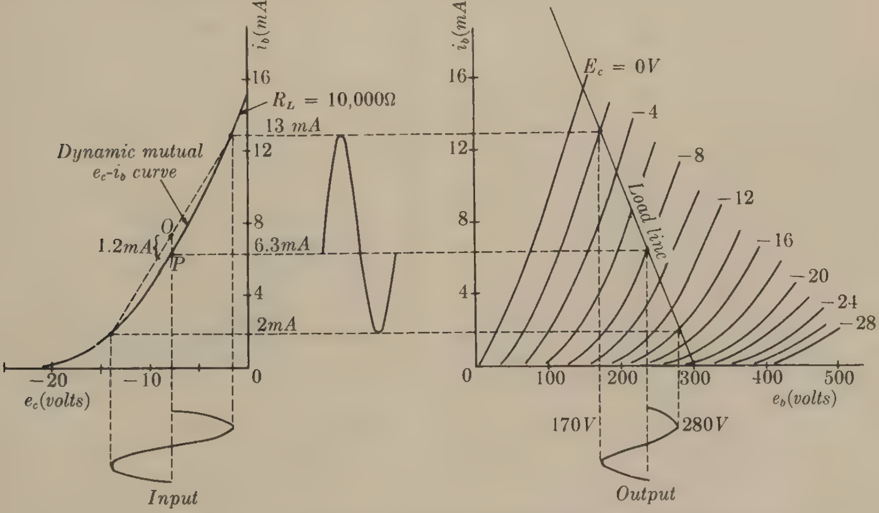


FIG.2-83 Operation of triode amplifier with load equal to anode resistance.

The peak value of the a-c component of anode voltage,

$$E_{a\ max} = 55 \text{ volts}$$

$\therefore$ the effective value of the a-c component of anode voltage,

$$E = \frac{55}{\sqrt{2}} \text{ volts}$$

Hence the a-c power delivered by the valve is

$$P_{a\text{-}c} = \frac{55}{\sqrt{2}} \times \frac{5.5}{\sqrt{2}} \times \frac{1}{1000}$$

$$= .154 \text{ watts.}$$

The percentage of second harmonic distortion may be determined from the graph in the following manner. If the amplifier were distortionless the dynamic $e_c - i_b$ curve would be linear over the limits of operation shown in Fig. 2-83. The operating point on this line corresponding to a grid bias of -8 volts is at the point O. The corresponding value of anode current is $7.5mA$. Then the percentage second harmonic distortion

$$= \frac{\text{current at } O - \text{current at } P}{\text{total current swing}} \times 100$$

$$= \frac{7.5 - 6.3}{11} \times 100$$

$$\doteqdot 11\%.$$

This value of distortion is considerably above the value of 5% which is usually taken as the maximum allowable distortion.

The power output and distortion produced by a load of twice the anode resistance of the valve are now determined from the graphs of Fig. 2-84. The larger value of load resistance gives rise to a dynamic $e_c - i_b$ curve which is less steep than that of Fig. 2-83. The anode current corresponding to the same grid signal as before, is now found to range from a minimum $1.3mA$ to a maximum of $8.3mA$. The corresponding values of anode voltage are 270 volts and 130 volts respectively.

$$I_{a\ max} = 3.5mA$$

$$I = \frac{3.5}{\sqrt{2}} mA$$

$$E_{a\ max} = 70 \text{ volts}$$

$$E = \frac{70}{\sqrt{2}} \text{ volts}$$

$$P = \frac{3.5}{\sqrt{2}} \times \frac{70}{\sqrt{2}} \times \frac{1}{1000}$$

$$= .123 \text{ watts}$$

That is, the power output is reduced from .154 watts with a load of 10,000Ω to .123 watts with a load of 20,000Ω.

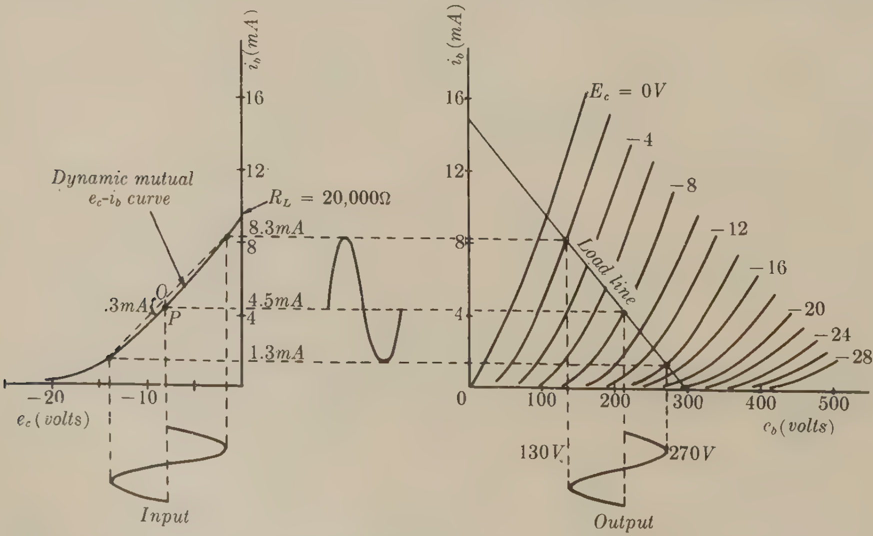


FIG. 2-84 Operation of triode amplifier with load equal to twice anode resistance.

The percentage second harmonic distortion

$$= \frac{\text{current at } O - \text{current at } P}{\text{total current swing}} \times 100$$

$$= \frac{4.8 - 4.5}{7} \times 100$$

$$= 4.3\%$$

By increasing the load from 10,000Ω to 20,000Ω, the distortion is reduced from 11% to 4.3%.

C. Matching a Loudspeaker to a Power Amplifier

Of particular importance in audio-frequency applications is the question of matching the impedance of a loudspeaker or group of loudspeakers to the impedance of the power amplifier. Loudspeakers operate at low voltage and high current. Since the impedance equals the effective voltage divided by the effective current, they are low-impedance devices. Similarly, valves are high-voltage, low-current, or high-impedance devices. To match a low-impedance loudspeaker to a high-impedance power amplifier stage for either maximum or optimum power output a matching transformer is used. The matching transformer takes power from the anode circuit at high voltage and low current and delivers it to the load at low voltage and high current.

The following example illustrates impedance matching for amplifiers.

A single-triode-valve power amplifier operating Class A, has an anode resistance $r_a = 800\Omega$. This amplifier is to be transformer coupled to a loudspeaker with a voice coil whose impedance $Z_L = 8\Omega$. The turns ratio of the transformer for an impedance match is given by the following expression.

$$\text{Turns ratio} = \sqrt{\frac{r_a}{Z_L}}$$

$$= \sqrt{\frac{800}{8}}$$

$$= 10$$

Therefore, the required turns ratio is 10.

For an optimum match, that is, to obtain maximum undistorted power output in the above example the turns ratio is given by the following expression.

$$\text{Turns ratio} = \sqrt{\frac{2r_a}{Z_L}}$$

$$= \sqrt{\frac{1000}{8}}$$

$$\doteqdot 14$$

Therefore the required turns ratio is 14.

The optimum anode load, which is 1600Ω in the above example, is usually specified by the manufacturer for a given valve. For a push-pull amplifier the manufacturer specifies the optimum anode-to-anode load resistor.

The following example illustrates how the proper turns ratio is calculated for an optimum match in such a case.

A Class A push-pull amplifier using beam power valves is to be coupled to a speaker with a voice coil with impedance $Z = 8\Omega$. The optimum anode-to-anode load resistance $R_{aa} = 4000\Omega$ as stated by the manufacturer.

The turns ratio of the transformer for an impedance match is

$$\text{turns ratio} = \sqrt{\frac{R_{aa}}{Z}}$$

$$\text{thus, the turns ratio} = \sqrt{\frac{4000}{8}}$$

$$\doteqdot 22.$$

Therefore, the turns ratio of the primary to the entire secondary winding is 22.

Sufficient information is available for selecting a suitable transformer if the primary and secondary impedances and the maximum power to be handled are known. If the available choice of transformers is limited it is usually best to select one which will present a higher rather than a lower than rated impedance to the valves. A variation of ±10% in the reflected impedance makes little difference in the output power and distortion. Usually an output transformer is provided with several secondary terminals to match various secondary load impedances to a given primary load impedance.

D. Matching Several Identical Loudspeakers to a Power Amplifier

When more than one speaker is to be connected to an amplifier the total load impedance may often be adjusted to give a close match to the amplifier output impedance by trying series and parallel arrangements and combinations of these. The parallel connection is generally preferable since the disconnection of one speaker from the circuit does not interrupt the operation of the remaining speakers, whereas with the series connection an open-circuited speaker disconnects the whole series group.

The possible connections for one to four 8Ω speakers to the output transformer of a power output stage are shown in Fig. 2-85. If the power required per speaker is 10*W*, the power in watts required from the amplifier is the number of speakers multiplied by 10, regardless of the type of connection. The effective impedance of a number of speakers in series is the sum of their individual impedances. The effective impedance of a number of identical speakers connected in parallel is equal to the impedance of one speaker divided by the number of speakers. In Fig. 2-85(f) the four speakers are connected in series-parallel and the effective impedance of the combination is equal to the impedance of one speaker. The various voltages and currents are found by using the formulas

$$P = I^2Z \text{ and } P = \frac{E^2}{Z}$$

and are shown on the diagrams. If the output transformer has secondary taps for any of the effective speaker impedances shown in the diagrams, the amplifier can be properly loaded without the use of auxiliary transformers by employing the corresponding wiring arrangement shown in the diagrams.

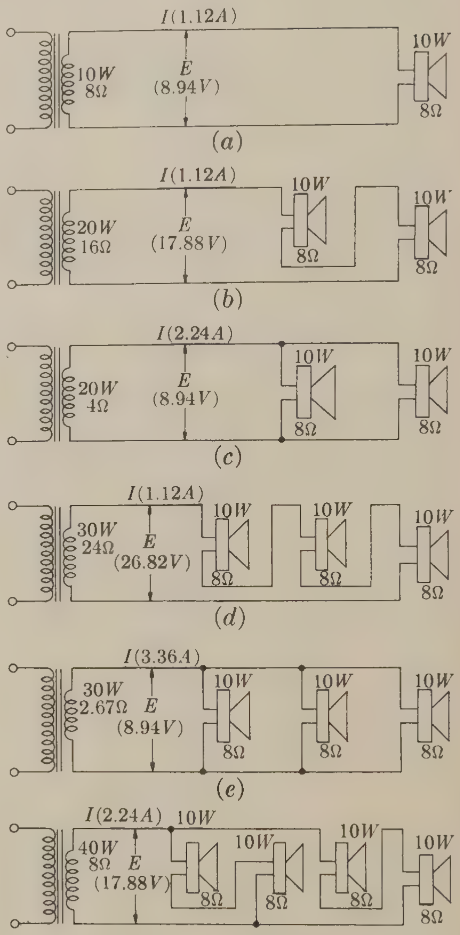


FIG. 2-85 Series and parallel connections to similar speakers.

E. Matching Several Dissimilar Loudspeakers to a Power Amplifier

When dissimilar speakers are to be connected to the same amplifier, the operating voltages must be determined in order that the system may be properly designed. For example, consider a system which requires two 16Ω speakers at 15*W* each and four 8Ω speakers at 5*W* each. Each group of speakers is connected in parallel (Fig. 2-86). The two 16Ω speakers have a combined impedance of 8Ω and require a total of 30*W* of power. From the formula $P = \frac{E_1^2}{Z}$ the voltage across

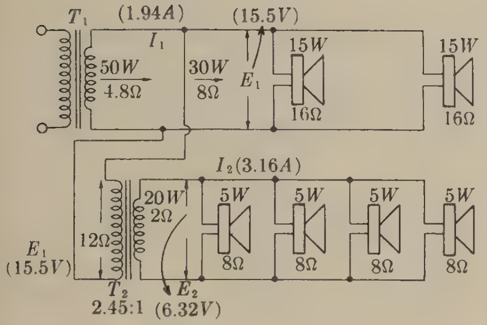


FIG.2-86 Matching dissimilar speakers to an amplifier.

these speakers is $E_1 = 15.5V$ and by Ohm's Law the total current $I_1 = 1.94A$. The four 8Ω speakers have a parallel impedance of 2Ω and require 20W. Thus, by using the same formulas, the voltage across these speakers is $E_2 = 6.32V$ and the total current $I_2 = 3.16A$.

If the secondary voltage of the output transformer T_1 is 15.5V, the 8Ω, 30W load may be connected directly to the transformer secondary. However the 2Ω, 20W load requires an auxiliary transformer T_2 which delivers 6.32V from an input of 15.5V. Its turns ratio from primary to secondary therefore equals

$$\frac{15.5}{6.32} = 2.45.$$

Transformers are usually specified by the impedances between which they operate. The secondary impedance is 2Ω and, assuming negligible power loss in the transformer, the primary impedance is

$$\frac{E_1^2}{P} \div \frac{240}{20} = 12.$$

The combined impedance of the two groups of speakers across the T_1 secondary is that of an 8Ω load in parallel with a 12Ω load and is 4.8Ω. A 4Ω or a 5Ω tap is sufficiently close. However, if these are not available and the amplifier has a given output impedance, say 500 ohms, a second auxiliary transformer may be inserted between the amplifier and the combined 4.8Ω load.

F. Matching Several Speakers to an Amplifier with a Fixed Output Impedance

In many cases the output transformer is part of the amplifier unit and requires a stated impedance across its secondary. The load impedances to be connected to the secondary are adjusted to present the stated impedance to the amplifier output terminals. For example, if four 8Ω speakers placed at intervals along a long line are to be connected to a 500Ω amplifier output the proper match may be obtained as follows. The combined impedance of the four loads in parallel must be 500Ω so that each speaker must present an impedance of 2000Ω. In order to do this an auxiliary transformer, 2000Ω-to-8Ω must be provided for each speaker as in Fig. 2-87. It is not desirable to have long lines from a transformer secondary to a low-impedance voice coil because of the power loss due to high current. Hence individual transformers are used. If the speakers are close together their voice coils may be connected in parallel giving a combined impedance of 2Ω which may be connected to the line through a 500Ω-to-2Ω transformer.

If the loudspeakers require different amounts of power, suitable power-dividing transformers must be used. Fig. 2-88 shows an amplifier with a 500Ω output impedance connected to three speakers which require 5W, 5W, and 10W respectively. The total power required is 20W

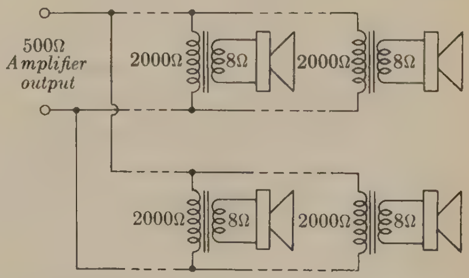


FIG.2-87 Matching to similar speakers at distant points.

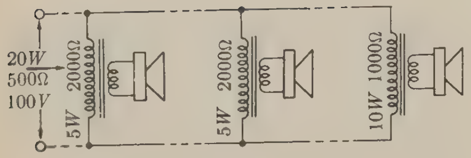


FIG. 2-88 Matching to speakers of different impedances and power requirements.

and thus the voltage across the output is $E = 100V$.

The impedance necessary for each load in order for it to absorb the required power when placed across the $100V$ is calculated as follows.

For each 5-watt load,

$$Z = \frac{E^2}{P} = \frac{100\times 100}{5} = 2000\Omega.$$

For the 10-watt load,

$$Z = \frac{E^2}{P} = \frac{100\times 100}{10} = 1000\Omega.$$

The effective impedance of the three loads in parallel is 500Ω.

Since the primary impedance for each transformer is now known, it is only necessary to make the secondary impedance correspond to that of the voice coil of the particular speaker. For example, if the $10W$ speaker has a 16Ω voice coil a 1000Ω-to-16Ω transformer is required.

2.21 Classification of amplifiers. The amplifiers described in the present chapter have been variously classified as to class of operation, type of service, kind of circuit and type of output. As there is considerable overlapping in these classifications, a summary is included in the following paragraphs.

A. *Class of Operation*

Amplifiers are classified according to the conditions under which they are made to operate. A Class A amplifier is one in which the grid bias and the alternating grid voltage are such that anode current flows at all times as shown in Fig. 2-89 (a).

In a Class B amplifier the grid bias is approximately equal to cut-off so that the anode current is approximately zero when no input voltage is applied. Hence the anode current flows for approximately one half of each cycle when an alternating grid voltage is applied. The operation of a Class B amplifier is shown in Fig. 2-89 (b).

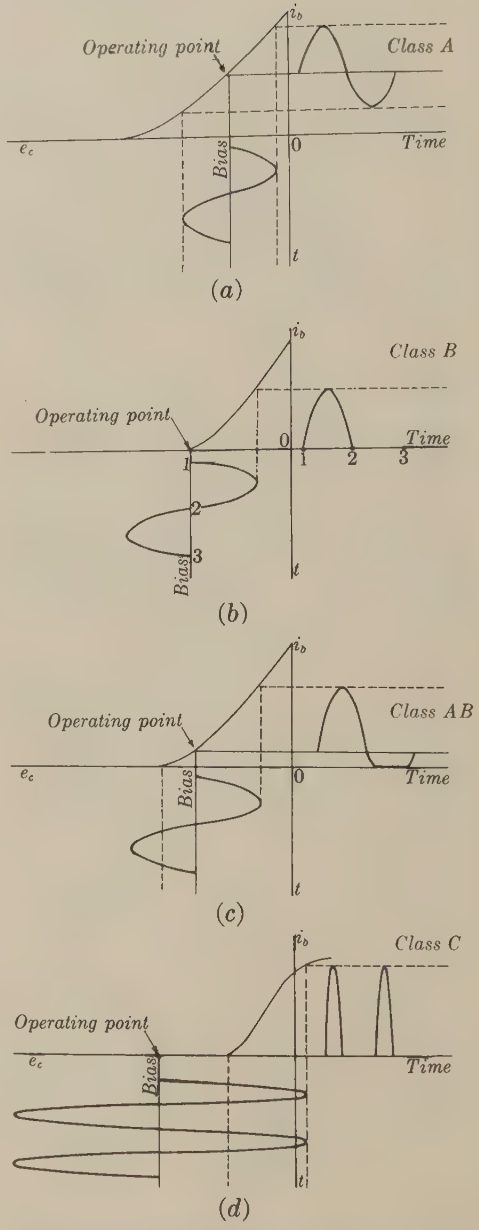


FIG. 2-89 Classification of amplifiers (a) Class A (b) Class AB (c) Class B (d) Class C.

A Class AB amplifier is one in which the grid bias and the alternating grid voltage are such that anode current flows for appreciably more than half but less than the entire alternating grid-voltage cycle as shown in Fig. 2-89 (c).

In a Class C amplifier the grid bias is appreciably greater than cut-off value so that the anode current is zero when no alternating grid voltage is applied. Thus the anode current flows for less than one half of each cycle when an alternating grid voltage is applied. Class C operation is shown in Fig. 2-89 (d).

The subscript 1 is added to the letter or letters of the class identification to indicate that grid current does not flow during any part of the input voltage cycle. The subscript 2 is added to indicate that grid current flows during part of the input voltage cycle, for example, class A_1, class AB_2.

B. Type of Service

Amplifiers are classified according to type of service as follows.

(i) Audio-frequency (a-f) amplifiers are used to amplify with constant gain at all frequencies within the audible range (15 to 15,000 c/s).

(ii) Radio-frequency (r-f) amplifiers are used to amplify at a single radio frequency or over a small band of radio frequencies.

(iii) Video-frequency (v-f) amplifiers are used to amplify with a uniform gain at frequencies extending over a very large range as required for television and radar (15 to 4,000,000 c/s).

(iv) Intermediate-frequency (i-f) amplifiers are r-f amplifiers used in special applications and the term does not refer to another class of amplifier.

C. Type of Circuit

Amplifiers are frequently used in cascade. That is, the output voltage of one stage is used as the input voltage of the succeeding stage. Thus amplifiers can be classified according to the type of circuit used to join the various stages. The usual types are resistance-capacitance, inductance-capacitance, and transformer coupling circuits.

D. Kind of Output

Amplifiers are classified as power or voltage amplifiers according to the output. Voltage amplifiers are designed to give a large voltage gain, power output being a minor consideration. Power amplifiers are used to deliver a large amount of power to a load, voltage amplification being incidental. Either voltage amplification or power amplification may be employed in any of the a-f, r-f, or v-f amplifier classes.

2.22 A public-address (P-A) system. The amplifier shown in Fig. 2-90 is typical of amplifiers used in the Service, in public address systems for hailing other ships and for recreational purposes within the ship itself. This amplifier contains the following components.

(i) Microphone
(ii) A-f voltage amplifier
(iii) Paraphase amplifier
(iv) Power amplifier
(v) Control circuits
(vi) Power supply.

A. The Microphone

The microphone is a dynamic unit fitted with a press-to-talk switch and a length of shielded 3-conductor cable. A description of this type of microphone is given in Chapter VIII.

B. The A-F Voltage Amplifier

The a-f voltage amplifier V_1 is a twin triode with anodes and grids respectively connected in parallel and is operated class A_1. The primary of the input transformer T_1 consists of two similar sections normally connected in parallel. R_1 is connected across the primary of T_1 and is the microphone-line-terminating re-

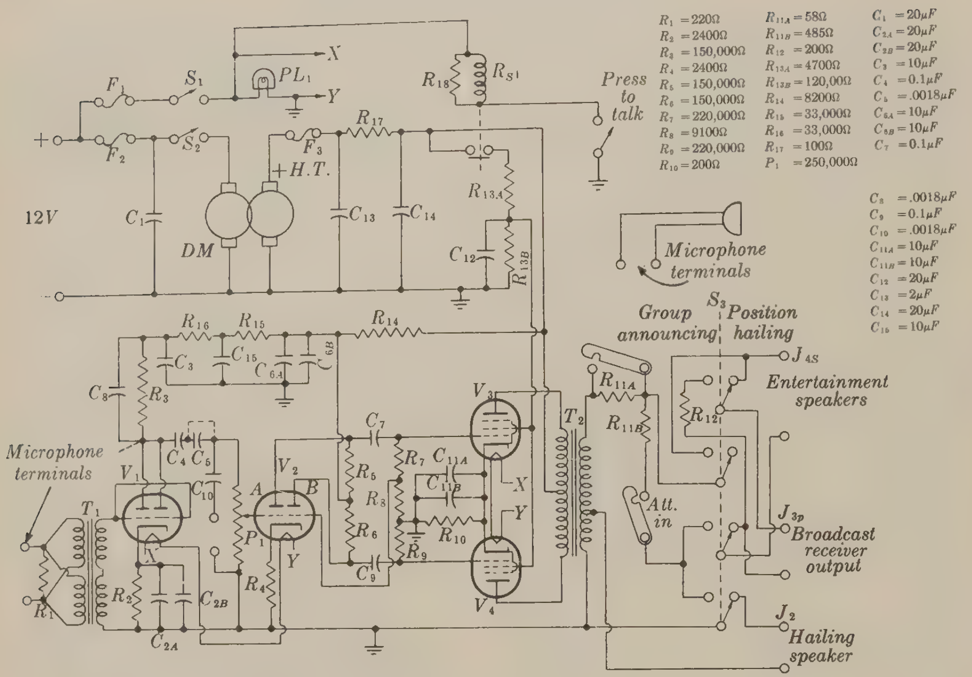


FIG.2-90 Typical circuit for public-address system.

sistor. Cathode bias is developed across R_2, by-passed by C_2. The output voltage of V_1 appearing across R_3, is coupled to the grid of the next valve through the resistance-capacitance network consisting of R_3, C_8, C_4, C_5, C_{10}, and P_1. This is a frequency-response adjusting network which is discussed later.

C. The Paraphase Amplifier

The paraphase amplifier V_2 is a twin triode, one section acting as an amplifier and the other as a phase inverter, both being operated class A_1. The cathode resistor R_4, is not by-passed and provides degeneration which improves the frequency response. P_1 is a volume control in the grid circuit of V_{2A}. The output of V_{2A} appears across the anode load resistor R_5 and is coupled through the blocking capacitor C_7 to the grid resistors, R_7 and R_8, of the power amplifier V_3. The portion of the output voltage which appears across R_8 is applied to the grid of V_{2B}. The output of this section appears across the anode load resistor R_6 and is coupled through C_9 to the grid resistor R_9 of V_4. From the values of resistance given in Fig. 2-90 it is seen that R_5 is equal to R_6 and that, within practical limits, R_7 together with R_8 equals R_9.

D. The Power Amplifier

The power amplifier stage consists of two beam power valves in a push-pull circuit and is operated class AB_1. Cathode bias is developed across R_{10}, by-passed by C_{11}. The power amplifier output is transformer coupled either to the hailing loudspeaker, via outlet jack J_2, or to a 200Ω transmission line feeding the voice coils of a bank of remotely-located entertainment loudspeakers, via outlet jack J_{4S}.

Load switching is accomplished by means of a 2-gang selector switch S_3. When this switch is in the group announcing position:

(i) the hailing loudspeaker is disconnected from the amplifier
(ii) the broadcast receiver is disconnected from the entertainment speakers
(iii) a dummy load R_{12} is connected to the broadcast receiver
(iv) the amplifier is connected to the entertainment speakers.

When this switch is in the hailing position:

(i) the amplifier is disconnected from the entertainment speakers
(ii) the broadcast receiver is disconnected from its dummy load and connected to the entertainment speakers
(iii) the hailing loudspeaker is connected to the amplifier.

An inverted-L type attenuator, R_{11A} and R_{11B}, which may be cut into, or out of, the 200Ω output circuit by metal straps, allows the normal output to be limited when the number of remote speakers is reduced. The secondary of the output transformer T_2 is tapped about $\frac{1}{4}$ way up from ground to match the 15Ω hailing loudspeaker.

E. Power Supply and Control Circuits

The four valve heaters are connected in series-parallel across the 12V battery supply through a filament switch S_1 and a 3A fuse F_1. The H.T. is obtained from a dynamotor DM, operated from the battery and controlled by a toggle switch S_2 in series with a 20A protecting fuse F_2. The H.T. is supplied through a $\frac{1}{4}A$ slo-blo type fuse F_3 and a π-section resistance-capacitance filter R_{17}, C_{13} and C_{14}, directly to the power stage. This voltage is further filtered, reduced, and decoupled by the H.T. voltage dropping resistor R_{14} and C_6 and applied to the paraphase amplifier. The voltage amplifier receives an H.T. which is still further filtered, reduced, and decoupled by the resistors R_{15} and R_{16} and the capacitors C_3 and C_{15}.

Keying of the H.T. circuit is effected by a relay RS_1 controlled by the press-to-talk switch on the microphone. The relay opens and closes the H.T. supply to the screen voltage divider R_{13A} and R_{13B} of the power stage. The coil of the screen-current-control relay and a pilot lamp PL_1 are connected in parallel across the heater circuit. The capacitor C_1 is a surge buffer.

The resistance-capacitance tone-control network, coupling the output of the voltage amplifier to the phase-inverter stage, may be adjusted to modify the normal frequency response of the amplifier to meet special operating conditions as shown in Fig. 2-91. Connecting links are used to provide the following arrangements.

(i) The arrangement of the coupling circuit shown in (a) is the usual resistance-capacitance network which gives a flat overall response.
(ii) In (b) the total coupling capacitance is reduced by putting C_4 and C_5 in series. This attenuates the low frequencies more than the high frequencies to cause the higher frequencies to predominate in the output.

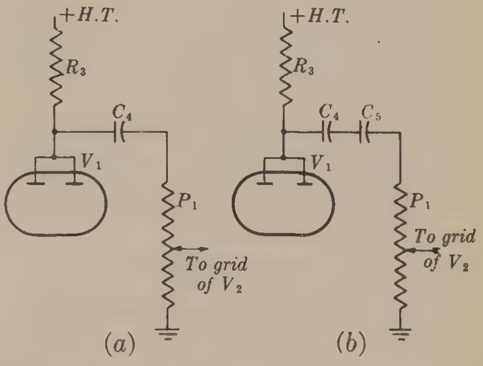


FIG. 2-91 Tone control network (a) Basic R-C coupling for flat overall response (b) Low-frequency attenuation.

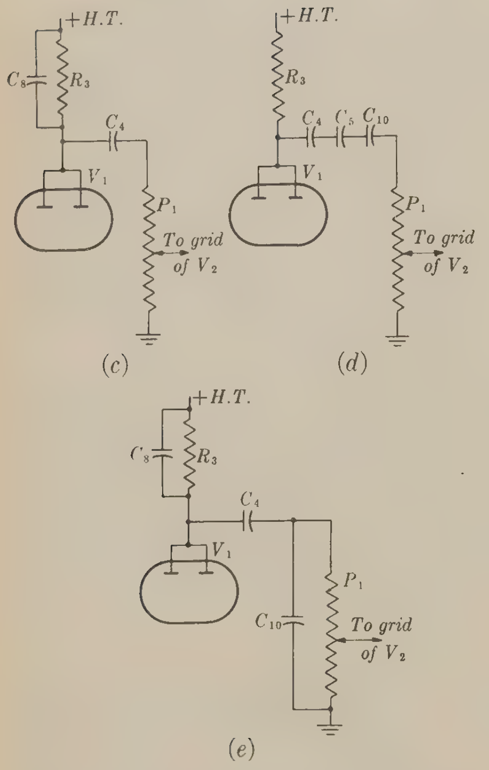


FIG.2-91 Tone control network (c) High-frequency attenuation (d) Additional low-frequency attenuation (e) Additional h-f attenuation.

(iii) In (c) C_8 is placed in parallel with R_3. This reduces the load at high frequencies more than at low frequencies to reduce the response at the high frequency end.

(iv) In (d) C_4, C_5 and C_{10} are connected in series to attenuate the low frequencies to a greater degree than in (b).

(v) In (e) C_8 is placed in parallel with R_3 and C_{10} with P_1 to attenuate the high frequencies to a greater degree than in (c).

2.23 Applied mathematics.

A. The Equivalent-Anode-Circuit Theorem and Gain of an Amplifier

The equivalent-anode-circuit theorem states that, as far as a-c components are concerned, the valve may be replaced in its circuit by an equivalent generator which has an internal resistance equal to the a-c resistance of the valve at the operating point chosen and which generates a voltage equal to the amplification factor of the valve multiplied by the grid-signal voltage. This is due to the fact that the grid is μ times as effective in controlling the anode current as the anode is. Thus the valve is represented as a source of constant voltage μe_g in magnitude. This constant-voltage generator has an internal resistance which is equal to the anode resistance r_a of the valve and is therefore in series with the generator as in Fig. 2-92 (a).

There is another form of equivalent circuit in which the valve is represented as a source of constant current $i = g_m e_g$ rather than a source of constant voltage. The anode resistance r_a of the valve is connected across the constant-current generator as in Fig. 2-92 (b). This type of equivalent circuit is more suitable for use with pentode valves where the anode resistance is high while the constant-voltage form is more suitable for use with triode valves.

An equation for the gain of an amplifier involving the constants of the valve and circuit can be derived as follows.

From the definition of the transconductance g_m of a valve (Chap. I)

$$\triangle i_b = g_m \triangle e_c \qquad (1)$$

where e_b is held constant.

Also from the definition of the anode resistance r_a

$$\triangle i_b = \frac{\triangle e_b}{r_a} \qquad (2)$$

where e_c is held constant.

Thus if both e_c and e_b are changed at the same time, the change in anode current is

$$\triangle i_b = g_m \triangle e_c + \frac{\triangle e_b}{r_a} \qquad (3)$$

In the case of an amplifier, the changes in current and voltage are the alternating components caused by the input volt-

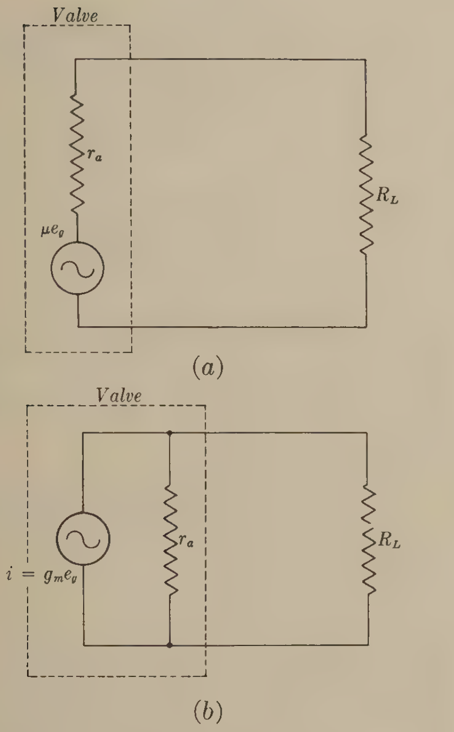


FIG. 2-92 Equivalent circuit of an amplifier (a) Constant voltage form (b) Constant current form.

age. The instantaneous value of an alternating voltage or current is the change in the voltage or current from the average value. Thus the changes in current and voltage in equation (3) can be replaced by the instantaneous alternating voltages and currents and the equation becomes

$$i_a = g_m e_g + \frac{e_a}{r_a} \qquad (4)$$

If the variations are kept within the linear portion of the dynamic $e_c - i_b$ curve g_m and r_a are constant. At any instant the total instantaneous anode voltage is

$$e_b = E_{bb} - i_a R_L \qquad (5)$$

That is, the change in e_b is $-i_a R_L$. But the change in e_b is e_a, the alternating component. Therefore

$$e_a = -i_a R_L \qquad (6)$$

Substituting for e_a in equation (4) results in

$$i_a = g_m e_g - \frac{i_a R_L}{r_a} \qquad (7)$$

and $$i_a\left(1 + \frac{R_L}{r_a}\right) = g_m e_g$$

Thus $$i_a = \frac{g_m e_g}{1 + \frac{R_L}{r_a}}$$

$$= \frac{r_a g_m e_g}{r_a + R_L}$$

$$= \frac{\mu e_g}{r_a + R_L} \qquad (8)$$

Therefore, $$\mu e_g = i_a(r_a + R_L) \qquad (9)$$

Equation (9) is the basic equation for the equivalent anode circuit using the constant-voltage generator μe_g. Since the same equation represents the action of an amplifier circuit as far as a-c components of current and voltage are concerned and also represents the action in a circuit of the type shown in Fig. 2-84 (a) this equivalent anode circuit represents the action of the amplifier for a-c components. Equation (9) can be derived by applying Ohm's Law to the circuit of Fig. 2-92 (a) in which the e.m.f. is μe_g, the current is i_a and the total resistance of the circuit is $r_a + R_L$. From equation (7)

$$g_m e_g = i_a\left(1 + \frac{R_L}{r_a}\right)$$

or $$g_m e_g = i_a R_L\left(\frac{1}{R_L} + \frac{1}{r_a}\right) \qquad (10)$$

Equation (10) is the basic equation for the constant-current generator form of the equivalent anode circuit. It may be derived by applying Ohm's law to the circuit of Fig. 2-92 (b) in which the generator current is $g_m e_g$, the voltage across the parallel resistance is $i_a R_L$ and the conductance is $\frac{1}{R_L} + \frac{1}{r_a}$.

The gain of the amplifier can be defined as

$$A = \frac{E_{a\,max}}{E_{g\,max}}$$

Equation (9) can be re-written for the

instant when e_g and i_a reach their maximum values by substituting

$$e_g = E_{g\,max}$$

and

$$i_a = I_{a\,max}$$

Thus equation (9) becomes

$$\mu E_{g\,max} = I_{a\,max}(r_a + R_L)$$

and

$$I_{a\,max} = \frac{\mu E_{g\,max}}{r_a + R_L} \tag{11}$$

But, considering magnitudes only,

$$E_{a\,max} = I_{a\,max} R_L \tag{12}$$

Substituting for $I_{a\,max}$ in equation (12) from equation (11) gives

$$E_{a\,max} = \frac{\mu E_{g\,max}}{r_a + R_L} \times R_L \tag{13}$$

which in turn is substituted into the gain equation to result in

$$\begin{aligned} A &= \frac{E_{a\,max}}{E_{g\,max}} \\ &= \frac{\frac{\mu E_{g\,max}}{r_a + R_L} \times R_L}{E_{g\,max}} \\ &= \frac{\mu R_L}{r_a + R_L}. \end{aligned} \tag{14}$$

B. *Decibels and Volume Units*

It is pointed out in section 2.6 that the ear has a logarithmic response to the intensity of sounds. For this reason it is convenient to choose units for the measurement of sound levels which also vary logarithmically. These units are the *bel* and the *decibel* and are defined by the formulas

$$\textit{\textbf{No. of bels}} = \log \frac{P_1}{P_2}$$

$$\textit{\textbf{No. of decibels}} = 10 \log \frac{P_1}{P_2}$$

where P_1 = larger of the two powers to be compared

P_2 = smaller power.

Example. A certain radio receiver uses a power valve V_1 in the final audio stage that delivers 4.5 watts to the loudspeaker. A second valve V_2 delivers 6.5 watts to the speaker. Express the increase in power as a percentage and in decibels.

$$\begin{aligned} \text{Percentage gain} &= \left(\frac{6.5-4.5}{4.5}\right) \times 100 \\ &\div 44\% \end{aligned}$$

$$\begin{aligned} \text{No. of decibels} &= 10 \log \frac{P_1}{P_2} \\ &= 10 \log \frac{6.5}{4.5} \\ &= 10 \times 0.1596 \\ &= 1.6 \end{aligned}$$

Thus, while the power has been increased by 44%, the increase in loudness is barely perceptible to the human ear since a gain of 1*db* is about the minimum which can be detected.

The bel and the decibel are only relative units and do not denote any definite amount of volume, power, current or voltage. In order to compare the outputs of two amplifiers without reference to their input powers, it is convenient to choose some arbitrary reference level. Although there is no single reference level which is used by all manufacturers, it is common practice to take a *power level of 6 milliwatts as zero db.* Any power level less than 6 milliwatts is indicated by *negative db* and any level above 6 milliwatts is indicated by *positive db.*

Example. An output transformer is rated for service at 36*db*. What is the power-handling capability of the transformer in watts?

$$\text{No. of decibels} = 10 \log \frac{P_1}{P_2}$$

$$P_2 = 0.006 \text{ watts}$$

Therefore

$$\begin{aligned} 36 &= 10 \log \frac{P_1}{0.006} \\ 3.6 &= \log P_1 - \log 0.006 \\ \log P_1 &= 3.6 + \bar{3}.7782 \\ &= 1.3782 \end{aligned}$$

Thus

$$P_1 = 23.9$$

The transformer is capable of handling 23.9 watts of power.

For measuring program levels on broadcast circuits transmitting speech and music, the industry has accepted a standard unit called the *volume unit* or *v-u*. The volume unit is based on a zero reference level of 1 milliwatt in a circuit of 600 ohms impedance. This measurement is made by a voltmeter of specified characteristics, called a *volume-level indicator*.

Example. An amplifier circuit produces a maximum output of 60 watts. Calculate its power rating in *db* and *v-u*.

$$\begin{aligned}\text{No. of } db &= 10 \log \frac{P_1}{P_2} \\ &= 10 \log \frac{60}{0.006} \\ &= 10 \log 10{,}000 \\ &= 40 \\ \text{No. of } v\text{-}u &= 10 \log \frac{60}{0.001} \\ &= 10 \log 60{,}000 \\ &= 47.7\end{aligned}$$

Formulas may also be developed for calculating power gain in terms of the input and output voltage or current as follows.

$$\begin{aligned}\text{No. of decibels} &= 10 \log \frac{P_1}{P_2} \\ \text{Substitute } P &= I^2R \\ \text{No. of decibels} &= 10 \log \frac{I_1^2R_1}{I_2^2R_2} \\ \text{No. of decibels} &= 10 \log \frac{I_1^2}{I_2^2} + 10 \log \frac{R_1}{R_2} \\ &= 20 \log \frac{I_1}{I_2} + 10 \log \frac{R_1}{R_2}\end{aligned}$$

If the input and output currents are made to flow through equal resistances,

$$R_1 = R_2$$

$$\therefore \textbf{No. of decibels} = 20 \log \frac{I_1}{I_2}$$

Also, $P = \dfrac{E^2}{R}$

$$\begin{aligned}\therefore \text{No. of decibels} &= 10 \log \frac{E_1^2}{R_1} \times \frac{R_2}{E_2^2} \\ &= 10 \log \frac{E_1^2}{E_2^2} + 10 \log \frac{R_2}{R_1} \\ &= 20 \log \frac{E_1}{E_2} + 10 \log \frac{R_2}{R_1}\end{aligned}$$

Again, if the input and output voltages are developed across equal resistances,

$$\textbf{No. of decibels} = 20 \log \frac{E_1}{E_2}$$

If the current and voltage are not in phase with each other, the power formula must include the input and output impedances and the phase angle. The power is given by the formula

$$P = EI \cos \theta$$

$$\begin{aligned}\therefore \text{No. of decibels} &= 10 \log \frac{P_1}{P_2} \\ &= 10 \log \frac{E_1I_1 \cos \theta_1}{E_2I_2 \cos \theta_2} \\ &= 10 \log \frac{E_1^2Z_2 \cos \theta_1}{E_2^2Z_1 \cos \theta_2}\end{aligned}$$

No. of decibels

$$= 20 \log \frac{E_1}{E_2} + 10 \log \frac{Z_2}{Z_1} + 10 \log \frac{\cos \theta_1}{\cos \theta_2}$$

where Z_1 and Z_2 are the impedances of the output and input circuits, θ_1 and θ_2 are the phase angles between current and voltage in the output and input circuits respectively.

A chart for converting current, voltage or power ratios into decibels is given in Appendix XVII. The table may also be used to convert *db* into current, voltage or power ratios.

C. *Phase Angle of an Amplifier*

It is shown in section 2.1 that the input voltage and the output voltage of an amplifier with a pure resistance load are 180° out of phase. When the load impedance is a combination of resistance and inductance the phase angle depends on the relative sizes of the load resistance, the load inductance, and the a-c resistance of the valve. This is seen in the vector diagram Fig. 2-93 (b) for the equivalent circuit of Fig. 2-93 (a).

The vector diagram is constructed on the I_a vector as a base. The p.d.'s across r_a and R_L are in phase with I_a since these are purely resistive components. The p.d. across L_L, I_aX_L,

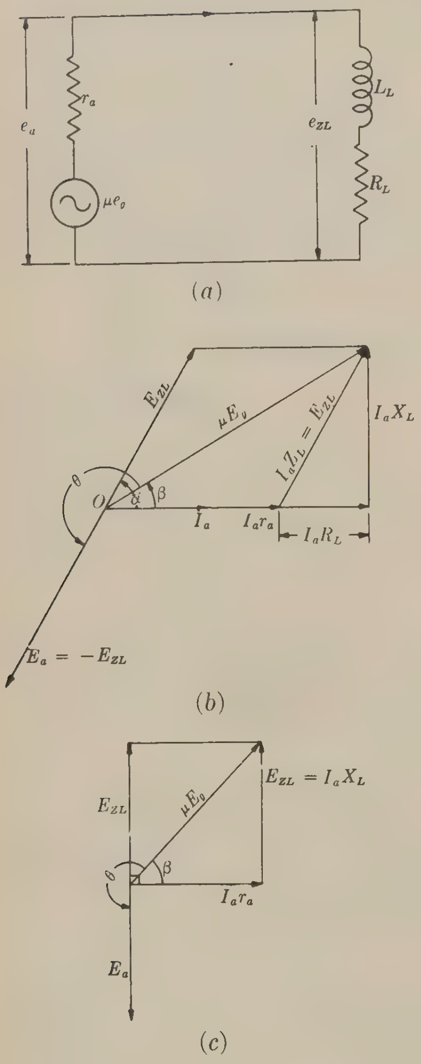


FIG.2-93 Phase shift in an impedance-loaded amplifier (a) Equivalent circuit diagram (b) and (c) Vector diagram.

leads I_a by 90° since L_L is a purely inductive component. Thus the total p.d. E_{ZL} across R_L and L_L, I_aZ_L, is the sum of the p.d.'s across R_L and L_L as shown. μE_g is the total voltage supplied to the circuit so that the sum of the p.d.'s across r_a and Z_L in the vector diagram will be equal to μE_g. The vector $I_aZ_L = E_{ZL}$ is re-plotted with its starting point at O. In the equivalent circuit, it is seen that $E_a = E_{ZL}$. Thus the E_a vector is drawn the same length as the E_{ZL} vector but 180° out of phase with it.

The angle by which E_{ZL} leads I_a is α. The angle by which E_a leads E_{ZL} is 180°. Thus E_a leads I_a by an angle of $(180°+\alpha)$. But μE_g and E_g lead I_a by the angle β. The phase angle of the amplifier is then $\theta = 180°+(\alpha-\beta)$. Since X_L changes with frequency it is seen on the vector diagram that the angle α, and therefore the phase angle of the amplifier θ, depend on the frequency of the input voltage and the size of R_L and L_L. Furthermore, the angle β depends on the size of α and r_a as well as on the above factors. Thus θ depends on the relative sizes of R_L, L_L, r_a and the frequency of the input voltage.

The following conclusions as to the dependence of the phase angle of the amplifier on the size of the components of the circuit and the frequency of the input voltage can be derived from the vector diagram.

(i) β increases as r_a decreases until $\beta = \alpha$ at $r_a = 0$. Therefore, the phase angle of the amplifier, $\theta = 180°+(\alpha-\beta)$, decreases as r_a decreases until at $r_a = 0$ in which case $\theta = 180°$.

(ii) If $X_L = 0$, that is the load is purely resistive, $\alpha = 0°$ and $\beta = 0°$. Therefore $\theta = 180°$ for all values of R_L.

(iii) If $R_L = 0$, that is the load is purely inductive, $\alpha = 90°$ and the vector diagram is as in Fig. 2-93 (c).

Thus
$$\begin{aligned}\theta &= 180°+(\alpha-\beta)\\ &= 180°+(90-\beta)\\ &= 270°-\beta.\end{aligned}$$

If, in addition $r_a = 0$, then $\beta = 90°$ and $\theta = 180°$.

When r_a increases, β decreases so that θ approaches 270° as r_a approaches infinity.

(iv) X_L increases as the frequency f of the input voltage increases. If $Z_L = R_L$, that is $L_L = 0$, then θ does not change with f. If L_L is not zero then α increases with f from $\alpha = 0°$ at $f = 0$ to approach $\alpha = 90°$ as X_L gets very large compared to R_L.

In the above discussions it is mentioned that certain conditions obtain in the amplifier circuit if $r_a = 0$. In practice all valves have an a-c resistance so that these conditions never actually exist. However, these conditions are approached by making Z_L very large compared to r_a. That is, α and β will be nearly equal when Z_L is very large.

D. Input Impedance for Resistive Load

In analysing the operation of amplifiers in cascade it is important to know what impedance an amplifier will appear to present to the output voltage of the preceding stage. The formula for the input impedance is developed using the circuit of Fig. 2-94. The interelectrode capacitances C_{ga} between grid and anode and C_{gk} between grid and cathode are drawn with dotted lines in their respective positions in the circuit. The anode-to-cathode capacitance has been included with the anode load impedance to give the load impedance Z_L which represents the total impedance of the anode circuit.

If this impedance, Z_L, is zero the input impedance as viewed from e_g is the reactance of C_{ga} and C_{gk} in parallel. That is, the input impedance is the reactance of the input capacitance C_i where

$$C_i = C_{gk} + C_{ga} \qquad (1)$$

If the impedance Z_L is not zero but purely resistive the anode voltage E_a is 180° out of phase with the grid voltage E_g. If the amplification of the stage is A then E_a is given by the relation

$$E_a = -AE_g$$

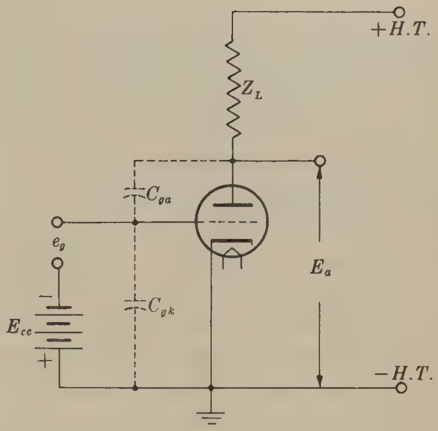


FIG.2-94 Input impedance of a triode amplifier.

Since the alternating voltage between grid and anode which appears across C_{ga} is the difference between E_g and E_a, therefore

$$\begin{aligned} E_g - E_a &= E_g - (-AE_g) \\ &= E_g + AE_g \\ &= E_g(1+A) \end{aligned}$$

That is, the alternating voltage across C_{ga} has been multiplied by a factor of $(1+A)$. The current I flowing through C_{ga} is from Ohm's Law,

$$\begin{aligned} I &= \frac{E_g(1+A)}{\dfrac{1}{\omega C_{ga}}} \\ &= \omega C_{ga} E_g (1+A). \end{aligned}$$

The current I_1 flowing through C_{gk} is from Ohm's Law,

$$I_1 = \frac{E_g}{\dfrac{1}{\omega C_{gk}}}.$$

I_g, the total current flowing in the input circuit, is equal to $I + I_1$.

Therefore

$$\begin{aligned} I_g &= \omega C_{ga} E_g (1+A) + \omega C_{gk} E_g \\ &= \omega E_g [C_{gk} + C_{ga}(1+A)] \qquad (2) \end{aligned}$$

Also, by Ohm's Law

$$\frac{E_g}{I_g} = \frac{1}{\omega C_i}$$

Thus $$C_i = \frac{I_g}{\omega E_g}$$

Thus, substituting from (2),

$$\begin{aligned} C_i &= \frac{\omega E_g [C_{gk} + C_{ga}(1+A)]}{\omega E_g} \\ &= C_{gk} + C_{ga}(1+A) \qquad (3) \end{aligned}$$

Thus it is seen that the input capacitance is greater when the amplifier has a load in the anode circuit than when it has no load. This is called *Miller Effect.*

When Z_L is reactive E_a and E_g are not exactly 180° out of phase and the current I_g does not lead E_g by exactly 90°. This indicates that there is an input resistance as well as an input capacitance. When the reactance of the load is capacitive the input resistance is positive. That is, power in the input circuit is lost to the anode circuit through C_{ga}. When the reactance of the load is inductive the input resistance is negative. A negative resistance is a component which supplies energy to the circuit instead of removing energy from the circuit. That is, the grid circuit receives energy other than that given to it by the E_g source. This energy comes from the anode circuit through the interelectrode capacitance C_{ga}. This phenomenon is called regeneration or positive feedback. The energy fed back helps E_g drive the current in the grid circuit, and hence the feedback is said to be positive.

E. Input Impedance for Reactive Loads

The quantities of charge corresponding to the r.m.s. values of grid voltage on the interelectrode capacitances for zero load impedance are

$$Q_{gk} = C_{gk}E_g$$

and

$$Q_{ga} = C_{ga}E_g.$$

The input capacitance may be found as follows. The sum of the charges caused by E_g is

$$Q_{gk}+Q_{ga} = C_{gk}E_g+C_{ga}E_g = E_g(C_{gk}+C_{ga}). \quad (4)$$

Because the capacitance of any capacitor is defined as

$$C = \frac{Q}{E},$$

the equivalent capacitance holding the charge $Q_{gk}+Q_{ga}$ is

$$C_i = \frac{Q_{gk}+Q_{ga}}{E_g} = C_{gk}+C_{ga}. \quad (5)$$

This is the input capacitance because it is the input voltage E_g which causes the alternating charge on the capacitances. A comparison of equations (5) and (1) shows that the same result is obtained by two different methods of development. The latter method is unwieldy in this simple example but can be used conveniently in the case where Z_L is not zero as follows.

Suppose Z_L is purely resistive. The charge on C_{gk} due to E_g is

$$Q_{gk} = C_{gk}E_g. \quad (6)$$

As was stated previously the alternating voltage between grid and anode which appears across C_{ga} is the difference between E_g and E_a.

Therefore,

$$\begin{aligned} E_g-E_a &= E_g-(-AE_g) \\ &= E_g+AE_g \\ &= E_g(1+A) \quad (7) \end{aligned}$$

and the charge on C_{ga} is

$$Q_{ga} = C_{ga}E_g(1+A). \quad (8)$$

The equivalent capacitance holding the total charge caused by applying E_g to the circuit is

$$\begin{aligned} C_i &= \frac{Q_{gk}+Q_{ga}}{E_g} \\ &= C_{gk}+C_{ga}(1+A). \quad (9) \end{aligned}$$

The input impedance in this case is the capacitive reactance of the input capacitance C_i.

In general when Z_L is not purely resistive, E_a is not equal to $-AE_g$ as in the above case but is equal to $-AE_g\cos\theta$ where θ is the angle by which the alternating voltage E_{Z_L} which is equal to $-E_a$, developed across Z_L, leads E_g (Fig. 2-95). The voltage across C_{ga} is now

$$E_g-(-AE_g\cos\theta) = E_g(1+A\cos\theta) \quad (10)$$

The charge on C_{gk} is

$$Q_{gk} = C_{gk}E_g \quad (11)$$

and the charge on C_{ga} is

$$Q_{ga} = C_{ga}E_g(1+A\cos\theta) \quad (12)$$

The equivalent capacitance holding the total charge is

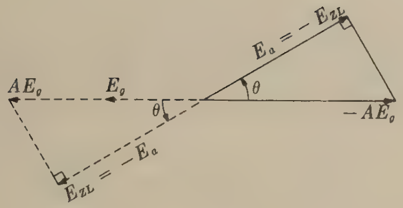


FIG.2-95 Vector diagram of input and output voltages for reactive load.

$$C_i = \frac{Q_{gk}+Q_{ga}}{E_g}$$
$$= C_{gk}+C_{ga}(1+A\cos\theta) \qquad (13)$$

It is shown in part *F* that in this case the input impedance has a resistive component known as the input resistance given by

$$R_i = -\frac{1}{A\omega C_{ga}\sin\theta} \qquad (14)$$

The input impedance consists of C_i and R_i in parallel. When Z_L has a capacitive component the voltage across Z_L lags E_g and the angle θ is negative. In this case sin θ is negative and R_i is positive. Since cos θ is positive when θ is either a negative or positive acute angle C_i is always positive. Thus Z_L is made up of positive resistance and capacitive reactance which indicates that energy is being fed into the anode circuit by way of C_{ga}.

When Z_L has an inductive component the angle θ is positive. In this case sin θ is positive and R_i is negative. This is the case of regeneration or positive feedback.

F. Alternative Method for Finding the Input Impedance

In general, the amplification of an amplifier stage is a complex quantity and is given by $A(\cos\theta+j\sin\theta)$ where A is the absolute ratio of the voltage developed across Z_L and the input voltage. Thus

$$A = \frac{E_a}{E_g}.$$

The angle θ is the phase angle between the two voltages.

It is desired to find the input impedance of the amplifier. The case where $\theta = 0°$ and E_g is such as to cause no grid current in the valve is considered first. This means that the current that flows in the grid circuit must divide and flow through the interelectrode capacitances C_{ga} and C_{gk}. The current I_1 through C_{gk} is given by

$$I_1 = E_g j\omega C_{gk}. \qquad (1)$$

The voltage across C_{ga} is the difference between E_g and E_a. Thus the current I_2 through C_{ga} is given by

$$\begin{aligned} I_2 &= (E_g - E_a)j\omega C_{ga} \\ &= (E_g + AE_g)j\omega C_{ga} \\ &= E_g(A+1)j\omega C_{ga}. \qquad (2) \end{aligned}$$

The total current is

$$\begin{aligned} I &= I_1 + I_2 \\ &= E_g j\omega C_{gk} + E_g(A+1)j\omega C_{ga}. \qquad (3) \end{aligned}$$

The input impedance of the amplifier is

$$Z_i = \frac{E_g}{I}. \qquad (4)$$

Substituting from (3) into (4)

$$\begin{aligned} Z_i &= \frac{E_g}{E_g j\omega C_{gk} + E_g(A+1)j\omega C_{ga}} \\ &= \frac{1}{j\omega[C_{gk}+(A+1)C_{ga}]}. \qquad (5) \end{aligned}$$

This indicates that the input impedance is purely capacitive and that the input capacitance is

$$C_i = C_{gk}+C_{ga}(1+A) \qquad (6)$$

For the general case where θ is not necessarily zero the input impedance from equation (5) becomes

$$\begin{aligned} Z_i &= \frac{1}{j\omega C_{gk}+[A(\cos\theta+j\sin\theta)+1]j\omega C_{ga}} \\ &= \frac{1}{j\omega C_{gk}+A\cos\theta j\omega C_{ga}+jA\sin\theta j\omega C_{ga}+j\omega C_{ga}} \\ &= \frac{1}{j\omega[C_{gk}+AC_{ga}\cos\theta+C_{ga}]-A\omega C_{ga}\sin\theta} \\ &= \frac{1}{j\omega[C_{gk}+C_{ga}(1+A\cos\theta)]-A\omega C_{ga}\sin\theta}. \qquad (7) \end{aligned}$$

For convenience the input admittance, Y_i, is used.

$$
\begin{aligned}
Y_i &= \frac{1}{Z_i}\\
&= j\omega[C_{gk}+C_{ga}(1+A\cos\theta)]-A\omega C_{ga}\sin\theta\\
&= -A\omega C_{ga}\sin\theta+j\omega[C_{gk}+C_{ga}(1+A\cos\theta)].
\end{aligned}
\qquad (8)
$$

The first term of equation (8) is the conductance and the second term is the susceptance. Note that the susceptance is capacitive. Thus the input resistance is $\dfrac{1}{A\omega C_{ga}\sin\theta}$ and the input capacitance is $C_{gk}+C_{ga}(1+A\cos\theta)$.

G. Theoretical Anode Efficiency

If the e_b-i_b curves are shown as straight lines as in Fig. 2-96 the theoretical anode efficiency may be found for a triode operated class A_1 with maximum power output and a given anode supply voltage. The load resistance R_L is equal to $2r_a$ and i_b is assumed to equal zero. The total input swing is from a to b about the operating point c. E_a is the effective value of the corresponding anode voltage and thus the peak value is $\sqrt{2}E_a$. Since $R_L = 2r_a$, $E_{b\,min}$ is $\frac{1}{3}E_{bb}$ and therefore $E_{b\,min}$ is $\frac{1}{2}E_{b0}$.

Then from the figure

$$\sqrt{2}E_a = \frac{E_{b0}}{2}$$

$$\sqrt{2}I_a = I_{b0}.$$

Since the load is a pure resistance $\cos\theta$ is unity and the a-c power output is

$$
\begin{aligned}
P_{a\text{-}c} &= E_aI_a\cos\theta\\
&= E_aI_a
\end{aligned}
$$

and the percentage anode efficiency η_a is given by

$$
\begin{aligned}
\eta_a &= \frac{E_aI_a}{E_{b0}I_{b0}}\times 100\\
&= \frac{\dfrac{E_{b0}}{2\sqrt{2}}\times\dfrac{I_{b0}}{\sqrt{2}}}{E_{b0}I_{b0}}\times 100\\
&= 25\%
\end{aligned}
$$

In practice this value is not obtained because of the curvature of the e_b-i_b curves.

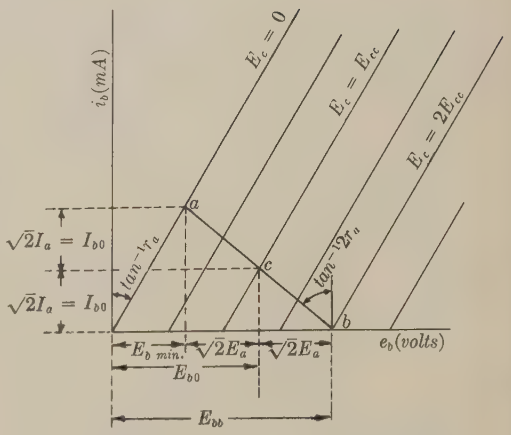


FIG. 2-96 Graph for calculating theoretical anode efficiency.

H. Percentage Second Harmonic Generated in a Valve Due to the Curvature of the Dynamic e_c-i_b Curve

The dynamic e_c-i_b curve is represented approximately by the quadratic equation

$$i_a = Ae_g+Be_g^2 \qquad (1)$$

where A and B are constants.

The load impedance is assumed to be a pure resistance and the grid-signal voltage e_g is assumed sinusoidal and is expressed as

$$e_g = \sqrt{2}E_g\cos\omega t. \qquad (2)$$

Substituting from (2) into (1)

$$i_a = A\sqrt{2}E_g\cos\omega t+B(\sqrt{2}E_g\cos\omega t)^2$$

Since $2\cos^2\omega t = 1+\cos 2\omega t$, then

$$
\begin{aligned}
i_g &= A\sqrt{2}E_g\cos\omega t+BE_g^2(1+\cos 2\omega t)\\
&= BE_g^2+A\sqrt{2}E_g\cos\omega t+BE_g^2\cos 2\omega t.
\end{aligned}
\qquad (3)
$$

Since this expression consists of a constant term as fundamental and a second harmonic the following substitutions are made. The constant term BE_g^2 represents the increase in the d-c component of anode current when a signal is applied and is represented by I_{ba}, the amplitude of the fundamental by $\sqrt{2}I_{a1}$ and the amplitude of the second harmonic by $\sqrt{2}I_{a2}$.

Thus

$i_a = I_{ba}+\sqrt{2}(I_{a1}$ [illegible] $t+I_{a2}\cos 2\omega t)$. (4)

The total anode current $i_b = i_a+I_{b0}$ where I_{b0} is the steady no-signal anode current. Since the anode current is a maximum when $\omega t = 0$ and a minimum when $\omega t = \pi$,

$$I_{b\,max} = I_{b0}+I_{ba}+\sqrt{2}I_{a1}+\sqrt{2}I_{a2} \qquad (5)$$

and

$$I_{b\,min} = I_{b0}+I_{ba}-\sqrt{2}I_{a1}+\sqrt{2}I_{a2} \qquad (6)$$

Thus by subtracting (6) from (5)

$$\sqrt{2}I_{a1} = \frac{I_{b\,max}-I_{b\,min}}{2}. \qquad (7)$$

By comparing (3) and (4) it may be seen that

$$\sqrt{2}I_{a2} = BE_g^2 = I_b$$

By adding (5) and (6) and substituting $I_{ba} = \sqrt{2}I_{a2}$

$$\sqrt{2}I_{a2} = I_{ba} = \frac{1}{2}\left(\frac{I_{b\,max}+I_{b\,min}}{2}-I_{b0}\right) \qquad (8)$$

The percentage second harmonic in the anode current and hence in the anode voltage is given by

$$\frac{\sqrt{2}I_{a2}}{\sqrt{2}I_{a1}}\times 100. \qquad (9)$$

By substituting from (7) and (8) this becomes

$$\frac{\frac{1}{2}(I_{b\,max}+I_{b\,min})-I_{b0}}{(I_{b\,max}-I_{b\,min})}\times 100. \qquad (10)$$

The values in this expression may be found from the load line on the e_b-i_b curves. The permissible upper limit of second harmonic generation is ordinarily 10%, but for certain applications this is only 5%.

I. Operation of a Push-Pull Class A Amplifier Beyond the Linear Region

If the grid-signal voltage e_g is sinusoidal and is expressed as

$$e_g = \sqrt{2}E_g\cos\omega t \qquad (1)$$

the anode current for no distortion is

$$i_b = I_{b0}+\sqrt{2}I_{a1}\cos\omega t. \qquad (2)$$

If distortion is present the anode current of the first valve may be expressed as the series

$$i_b = I_{b0}+I_{ba}+\sqrt{2}(I_{a1}\cos\omega t+ I_{a2}\cos 2\omega t+I_{a3}\cos 3\omega t+\dots) \qquad (3)$$

Thus an increase in average anode current when the signal is applied indicates that distortion is present. If the push-pull circuit is symmetrical the anode current of the second valve is a similar expression with ωt replaced by $\omega t+\pi$.

Thus

$$i_b' = I_{b0}+I_{ba}+\sqrt{2}[I_{a1}\cos(\omega t+\pi)+I_{a2}\cos(2\omega t+2\pi)+I_{a3}\cos(3\omega t+3\pi)+\dots] \qquad (4)$$

$$= I_{b0}+I_{ba}+\sqrt{2}\,[-I_{a1}\cos\omega t+I_{a2}\cos 2\omega t-I_{a3}\cos 3\omega t+\dots] \qquad (5)$$

If the output transformer is ideal the current ratio is the inverse of the turns ratio.

That is,

$$\frac{N_1}{N_2}=\frac{i_2}{i_b-i_b'} \qquad (6)$$

where

$\frac{N_1}{N_2}$ = turns ratio of half the primary to the whole secondary

i_2 = secondary current

i_b-i_b' = effective primary current, since the anode currents flow through the primary in opposite directions.

Substituting from (3) and (5) in equation (6) the secondary current is given as

$$i_2=2\frac{N_1}{N_2}\sqrt{2}(I_{a1}\cos\omega t+I_{a3}\cos 3\omega t+\dots) \qquad (7)$$

Thus the average components, the second harmonic and all other even harmonics generated within the valves are cancelled in the output.

The current through the H.T. supply is the sum of the currents through the two valves and is given by

$$i_b+i_b' = 2I_{b0}+2I_a+ \sqrt{2}\,(2I_{a2}\cos 2\omega t+2I_{a4}\cos 4\omega t+\dots) \qquad (8)$$

Thus in the H.T. supply only the even harmonic components are present. An increase of $2I_{ba}$ in the average anode current occurs when harmonics are being generated in the valves.

2.24 Questions and problems.

1. Draw a simple voltage amplifier circuit and label its components.
2. Explain with the use of graphs the variations in anode current and anode voltage
   (a) when no signal is applied to the grid
   (b) when a sinusoidal grid signal is applied.
3. What is the purpose of grid bias?
4. How is the value of grid bias chosen in designing an amplifier?
5. State three methods of obtaining bias and name the components used to obtain the bias in each case.
6. Calculate the size of the cathode resistor required to provide a bias of $-10V$ with an average current of $4mA$.
7. Why must the bias remain steady during operation of the amplifier? What component maintains the cathode at a steady value?
8. Explain the operation of the cathode by-pass capacitor
   (a) on the basis of charge and voltage
   (b) on the basis of current components.
9. Name two types of grid-leak bias.
10. Make a diagram of an amplifier to show each of the two types of grid-leak bias.
11. Explain with the aid of graphs
    (a) how grid-leak bias is developed
    (b) how the bias is maintained.
12. What happens to the value of grid-leak bias
    (a) if the amplitude of the signal increases
    (b) if the grid signal is removed?
13. Differentiate between static characteristics and dynamic characteristics of an amplifier.
14. Explain how the dynamic $e_c - i_b$ characteristic of an amplifier is obtained.
15. How is the position of the $e_c - i_b$ curve altered
    (a) if the anode load is increased
    (b) if the supply voltage is increased?
16. Show how the variations in anode current are determined from the $e_c - i_b$ curve when a sinusoidal input is applied to the grid.
17. What is meant by the term load line?
18. Explain what measurements would be taken in a circuit to plot a load line.
19. How is the load line obtained from a set of static $e_b - i_b$ curves?
20. Show how the graph of question 16 may be extended to determine the variations in output voltage from the load line.
21. Indicate how the variations in grid voltage, anode current and anode voltage may be plotted on a single graph using the load line.
22. Select from the receiving valve manual a triode valve having an amplification factor of about 20. From the $e_b - i_b$ curves given in the manual, plot a load line for a load of 25,000Ω and a supply voltage of 250 volts. Determine the value of cathode resistor required to establish a bias of -6 volts. Plot the variations in grid voltage, anode current and anode voltage corresponding to a sinusoidal input of 4 volts amplitude. Determine the amplitude of the output signal at the anode. What is the voltage gain?

23. Draw the equivalent circuit of a triode amplifier with a resistance load. Of what use is the equivalent circuit?

24. What is the formula for the gain of a triode amplifier with a resistance load?

25. Using this formula, calculate the gain of the circuit described in problem 22.

26. How may the gain of such an amplifier be increased?

27. Name three types of inter-stage coupling in voltage amplifiers.

28. Make a diagram of a two-stage resistance-capacitance coupled amplifier and label the components.

29. (a) What is the purpose of the coupling capacitor?
    (b) What is the purpose of the grid-leak resistor?

30. What factors determine the size of the coupling capacitor and grid-leak resistor for an a-f voltage amplifier?

31. State one advantage and one disadvantage of inductance-capacitance coupling.

32. How is the impedance of an anode load inductor calculated?

33. What is the impedance of an anode load inductor having a resistance of 500 ohms and an inductance of 20 henries when operated at a frequency of 200 cycles per second?

34. On what factors does the gain of an inductance-loaded amplifier depend?

35. Make a diagram of a transformer-coupled voltage amplifier.

36. What type of valve is usually employed in transformer-coupled voltage amplifiers?

37. (a) What is a satisfactory turns ratio for such a transformer?
    (b) Why can a higher turns ratio not ordinarily be used?

38. What is meant by the term frequency response of an amplifier?

39. How is a frequency-response curve determined?

40. Explain the meaning of the terms mid-frequencies, low frequencies, high frequencies when applied to an a-f voltage amplifier.

41. What factors determine the limit of the low-frequency response? Of the high-frequency response?

42. On what factors does the input capacitance of a resistance-loaded amplifier depend?

43. What effect does an increase in gain have on the input capacitance?

44. By reference to the valve manual, calculate the input capacitance of the triode valve selected in question 22 when operating with a gain of 18.

45. What effect would an increase in input capacitance produce in the shape of the response curve?

46. Make a diagram of a resistance-capacitance coupled pentode amplifier.

47. What is the purpose of
    (a) the screen dropping resistor
    (b) the screen by-pass capacitor?

48. How does such a circuit compare with the triode amplifier as to
    (a) size of anode load resistor
    (b) amplification?

49. How does the input capacitance of a pentode amplifier compare with that of a triode? Explain.

50. Name three types of distortion common in voltage amplifiers.

51. Explain with the use of the dynamic $e_c - i_b$ characteristic how distortion is caused by non-linearity of the characteristic.
52. In what other ways may amplitude distortion be introduced?
53. Explain how the flow of grid current introduces distortion in the anode circuit.
54. Why is amplitude distortion sometimes called harmonic distortion?
55. What components in an a-f amplifier other than the valve may introduce distortion?
56. Explain what is meant by frequency distortion.
57. What effect is produced on sound reproduction
    (a) by reducing the high-frequency components
    (b) by reducing the low-frequency components?
58. What components of the circuit are usually responsible for introducing frequency distortion?
59. What types of load circuits introduce phase distortion?
60. Why is phase distortion usually negligible in a-f amplifiers?
61. What is the purpose of an a-f power amplifier?
62. How do power amplifier valves differ from voltage amplifier valves?
63. In what way does the impedance of the load circuit affect the power output of the valve?
64. What formula is used in computing the power output of an amplifier?
65. What formula is used in computing the maximum power output?
66. Why is a power amplifier seldom operated with maximum output power?
67. How may distortion be reduced in a power amplifier circuit?
68. How is optimum power output obtained from a power amplifier?
69. How does optimum power compare with maximum power as to magnitude and distortion present?
70. What is meant by the power sensitivity of an amplifier? State the formula for computing the power sensitivity.
71. What is meant by the term anode efficiency of a power amplifier?
72. What is meant by anode dissipation?
73. How may the anode efficiency be increased?
74. What is the maximum anode efficiency possible with Class A power amplifiers?
75. How does the anode dissipation of a valve change when a signal is applied to the grid?
76. Account for the fact that a power amplifier valve may burn out if the speaker is disconnected during operation.
77. Make a circuit diagram to show two valves operating in parallel.
78. Compare the operation of two valves in parallel with that of a single valve as to
    (a) amplitude of grid signal permissible
    (b) power output
    (c) load impedance required for optimum output.
79. Make a circuit diagram of a push-pull a-f power amplifier and label the components.
80. State two advantages of push-pull operation over parallel operation.

81. Explain how currents in opposite directions in the primary of the output transformer produce currents that add together in the secondary winding.

82. Explain with the aid of graphs how distortion is reduced by operating valves in push-pull.

83. Why may each valve in a push-pull circuit be made to deliver twice as much power as the same valve in a single-ended amplifier?

84. Differentiate by means of graphs between Class A, Class AB and Class B operation of valves in push-pull.

85. Make a circuit diagram of a typical r-f voltage amplifier. Why are pentode valves used in r-f amplifiers?

86. What type of coupling is generally employed in r-f voltage amplifiers?

87. What advantage does tuned-transformer loading have over
    (a) resistance loading
    (b) inductance loading for r-f amplifiers?

88. What is the formula for the gain of an r-f amplifier with a tuned anode load?

89. In what respects does the equivalent circuit of the r-f pentode amplifier differ from that of the a-f triode amplifier?

90. What is meant by band-pass coupling?

91. State one type of circuit which employs band-pass coupling.

92. Explain why communications receivers usually have narrower band-pass characteristics than broadcast receivers.

93. (a) What use is made of radio-frequency power amplifiers?
    (b) What types of valves are employed for this purpose?

94. Explain using graphs the operation of a Class C r-f power amplifier. Indicate the portion of the cycle during which
    (a) grid current flows
    (b) anode current flows.

95. Explain how the pulses of current through the valve maintain a sinusoidal output from the load circuit.

96. (a) What is a driver stage?
    (b) Why are drivers required in high-power transmitters?

97. What advantage does a direct-coupled amplifier have over one with capacitance coupling?

98. Why can the coupling capacitor between stages of an *R-C* coupled amplifier not be short-circuited to provide direct coupling?

99. Draw a typical direct-coupled amplifier and explain its operation
    (a) when no signal is applied to the grid
    (b) when the grid is made more negative
    (c) when the grid is made less negative.

100. How does the cathode-loaded amplifier differ from the anode-loaded amplifier in construction?

101. Compare the cathode- and anode-loaded amplifiers as to gain and phase relationships between input and output.

102. (a) How does the cathode follower differ in construction from the cathode-loaded amplifier?
    (b) Make a diagram of each circuit and indicate the input and output terminals of each.

103. What is the significance of the name cathode follower?

104. (a) What is the maximum gain of a cathode follower?
    (b) Of what use is such a circuit?

105. How do the input and output impedances of a cathode follower compare with those of an amplifier?

106. What is the advantage of
    (a) a high input impedance
    (b) a low output impedance in an electrical circuit?

107. What is the chief characteristic of a video amplifier that makes it different from an a-f or r-f amplifier?

108. Why are wide band-pass amplifiers given the name video amplifiers?

109. Describe the harmonic composition of a square wave.

110. Why are video amplifiers required to amplify square waves and saw-tooth waves of voltage?

111. What change would result in the waveform of a square wave of voltage when amplified by a circuit with poor high-frequency response?

112. How must the phase of the output voltage compare with that of the input voltage throughout the video frequency range?

113. (a) What is meant by high-frequency compensation in a video amplifier?
    (b) How may this compensation be achieved?

114. Describe a method of obtaining low-frequency compensation.

115. (a) Why are pentode valves used in preference to triodes as video amplifiers?
    (b) How does a television pentode differ from an r-f pentode?

116. In what types of circuits are video power amplifiers used?

117. What types of valves are suitable for video power amplifiers?

118. State two methods of matching a video amplifier to a low-impedance circuit.

119. What is a phase-splitting circuit?

120. Draw a simple phase-splitting circuit and suggest an application for it.

121. Make a diagram of a paraphase amplifier and explain briefly how equal outputs are obtained from the two valves.

122. What advantage has a floating paraphase amplifier over the more conventional type?

123. What is meant by inverse feedback in an amplifier?

124. What is an alternative name for inverse feedback?

125. What is the purpose of introducing degeneration into a circuit?

126. (a) How is voltage degeneration obtained in a single-valve amplifier?
    (b) Describe briefly the operation of the circuit.

127. (a) How is current degeneration introduced into an amplifier?
    (b) Describe the operation of the circuit with degeneration.

128. On a dynamic $e_c - i_b$ characteristic, indicate the operating point and amplitude of signal required for Class A, Class AB, Class B and Class C operation respectively.

129. State one type of circuit for which each class of operation is suitable.

CHAPTER III

POWER SUPPLIES

3:1 General outline. Electrical energy is required to operate all radio equipment. The original sources of this energy are called the *primary sources*. These sources utilize either chemical or mechanical energy and convert it into electrical energy. Examples of primary sources are batteries, composed of either primary or secondary cells, d-c generators and a-c generators. The power supplied by these sources to the radio equipment is needed at two main voltage levels which are designated by the terms *low tension* (*L.T.*) meaning low voltage and *high tension* (*H.T.*) meaning high voltage. The L.T. power supply is mainly used to heat the cathodes of the valves while the H.T. power supply is used to apply d-c voltages to the anodes.

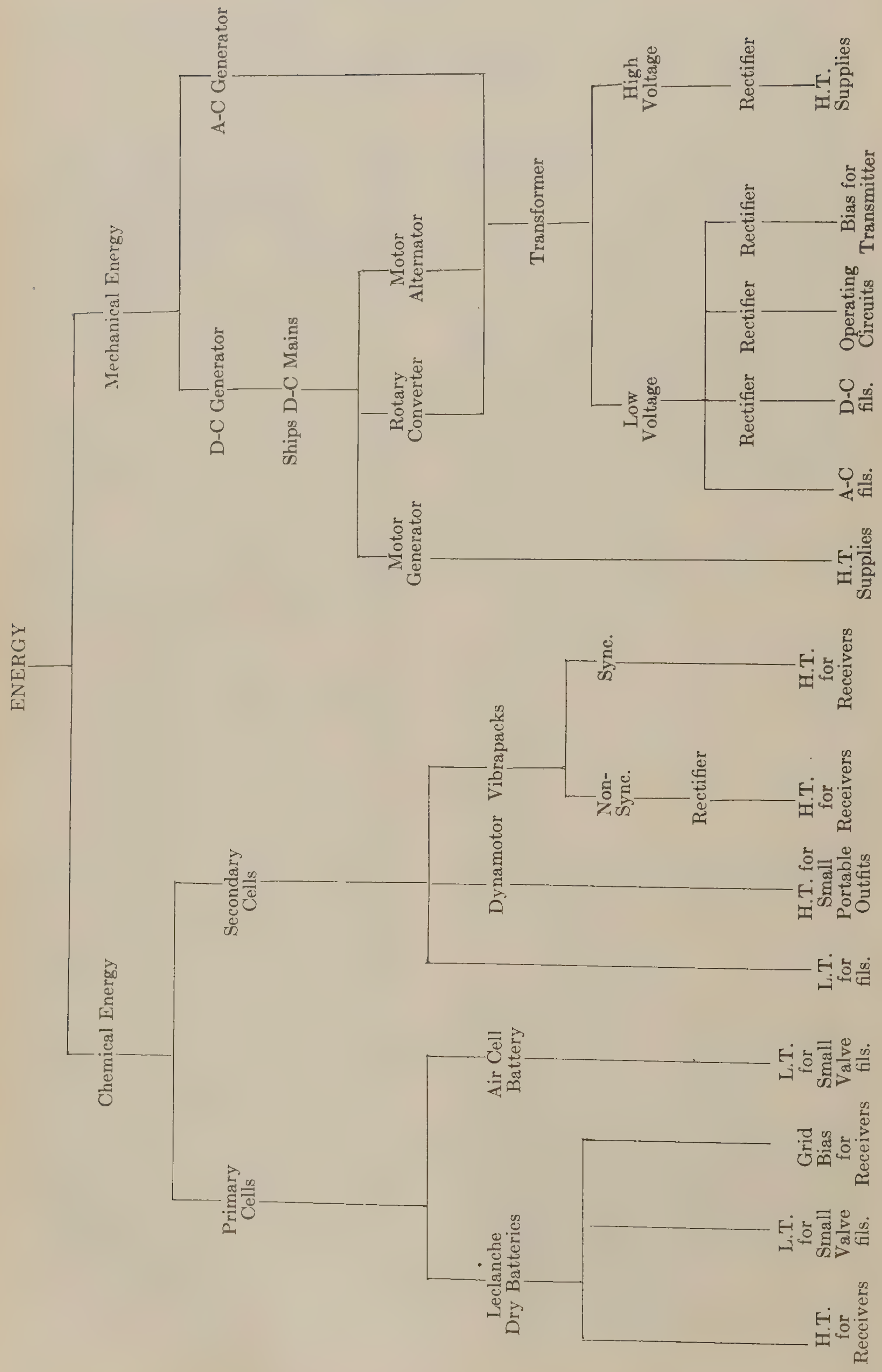
The above supply voltages may be directly available from the primary sources. In most cases, however, it is necessary to change the voltage from the primary source to one of different magnitude or type or both before it is suitable for use in the radio equipment. Most of the methods commonly used for supplying the proper L.T. and H.T. voltages are summarized in the table on page 3:2.

For equipments having a small power drain, for example small portable equipment, the H.T. is supplied from B batteries consisting of a number of dry cells in series and the L.T. is also supplied from primary cells. For portable equipment requiring a source which will deliver more power, secondary cells or batteries are useful. The voltage of a storage battery is suitable for L.T. directly but it is impractical to provide a sufficiently high voltage for H.T. by means of secondary cells alone because of the cumbersome bulk of a large number of cells connected in series. Consequently a dynamotor or a vibrapack is used to convert the low battery voltage to a suitable high voltage.

In most non-portable equipment the power is originally obtained from a-c or d-c generators. When d-c mains are used the voltage supplied is seldom at the proper level. The easiest way to change voltage level is to use a transformer. This necessitates changing the d-c to a-c before the voltage is stepped up or down to the desired values. The change from d-c to a-c may be made by means of a rotary converter or a motor alternator. After the d-c has been changed to a-c, the correct voltage for both the L.T. and H.T. is obtained by means of different windings on a single transformer. Since the H.T. must be a d-c voltage, the high-voltage a-c is reconverted to d-c by means of a rectifier circuit.

When a-c mains are used, the desired voltages are obtained from a transformer and the high voltage rectified as before.

Although the phrase *the power supply*, as applied to radio equipment includes, in its widest meaning, all the equipment from the primary source until the voltage is applied to the valves, it has taken on a more restricted meaning in referring only to the equipment which is installed in the radio set itself and is the last stage in providing the correct voltage for the valves. For example, in the case of the low-power portable equipment *the power supply* consists of the primary cells; but in the case in which the voltage goes


ENERGY
Mechanical Energy
Chemical Energy
A-C Generator
D-C Generator
Ships D-C Mains
Motor Alternator
Rotary Converter
Motor Generator
Transformer
High Voltage
Low Voltage
Rectifier
H.T. Supplies
Rectifier
Bias for Transmitter
Rectifier
Operating Circuits
Rectifier
D-C fils.
A-C fils.
H.T. Supplies
Secondary Cells
Primary Cells
Vibrapacks
Sync.
H.T. for Receivers
Non-Sync.
Rectifier
H.T. for Receivers
Dynamotor
H.T. for Small Portable Outfits
L.T. for fils.
Air Cell Battery
L.T. for Small Valve fils.
Leclanche Dry Batteries
Grid Bias for Receivers
L.T. for Small Valve fils.
H.T. for Receivers

through the sequence, d-c mains to rotary converter to transformer and rectifier, it is only the equipment from the transformer on that is commonly designated by the term *the power supply*.

3.2 Half-wave and full-wave rectification.

A. Rectifiers

A *rectifier* is a device which is used to change an a-c current into a d-c current. Any device that conducts current better in one direction than in the opposite direction can be used as a rectifier. There are several types of rectifiers such as diode valves (either high-vacuum or gas-filled), metallic rectifiers, crystal rectifiers, electrolytic rectifiers and mechanical rectifiers. Of these, only the diode valves and the metallic rectifiers are used in present-day power supplies. The hard or high-vacuum valve rectifier finds application in equipment where a small amount of power is needed or where an exceptionally high voltage is rectified. The soft or gas-filled valves are used when the H.T. power supply must be able to deliver larger amounts of power. Metallic rectifiers sometimes replace the hard-valve rectifiers in H.T. power supplies and are often used to provide a low-voltage d-c for relays.

B. Half-Wave Rectification

In a diode valve, current flows when the anode is positive with respect to the cathode. If an a-c voltage is applied across a diode valve, the above condition is true only during the positive half-cycles of the input a-c voltage. Therefore the a-c is changed into a pulsating direct current and rectification takes place. Since current flows only during one half of each input cycle, this type of rectification is called *half-wave rectification*.

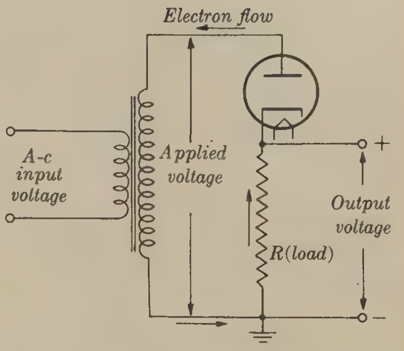


FIG.3-1 A simple half-wave rectifier circuit.

A simple half-wave rectifier circuit is shown in Fig. 3-1. The secondary of a transformer is connected across a diode and a load resistor R in series. The a-c voltage from the transformer is called the *applied voltage*. At the beginning of a positive half cycle, some electrons are attracted from cathode to anode in the valve. Consequently a small current flows from ground through R, through the valve and back to ground through the secondary of the transformer. As the anode becomes more positive, the current increases until the peak of the positive half cycle is reached. As the voltage decreases, the current also decreases, reaching zero when the applied voltage is again zero. This current produces a positive pulse of voltage across R. During a negative half cycle, the current is zero since electrons cannot flow from anode to cathode. The output voltage is therefore a series of pulses, all in the same direction, occurring during every positive half cycle in the input a-c.

The applied voltage, the current in the circuit and the output voltage across the load are shown graphically in Fig. 3-2.

C. Full-Wave Rectification

Full-wave rectification takes place when two or more rectifying elements are so arranged that a current flows through the load resistor in the same direction during both halves of an input a-c cycle.

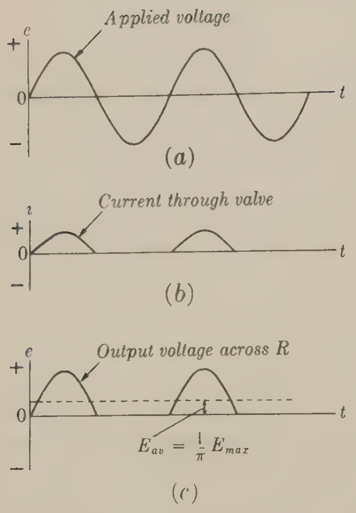


FIG.3-2 Waveforms of voltage and current in a half-wave rectifier.

The simplest form of a full-wave rectifier circuit is shown in Fig. 3-3 (a) and (b). The cathodes of two diodes are joined and the load resistor is connected at one end to the cathodes and at the other end to the centre of the transformer secondary winding. The anodes of the valves are connected one to each end of the secondary winding. The centre tap, marked C on the diagram, is usually grounded and is taken as the reference point when discussing potentials at other points in the circuit.

Fig. 3-3 (a) shows the conditions in the circuit during a positive half cycle of the input voltage and Fig. 3-3 (b) shows the conditions during a negative half cycle. During the positive half cycle 1, the end of the transformer marked A is positive with respect to the end marked B. Since point C is between A and B, the anode of V_1 is positive with respect to C and the anode of V_2 is negative. Consequently V_1 conducts and current flows from C through R through V_1 and back to C through the half of the secondary winding marked AC. This pulse of current through R causes a pulse of positive voltage to appear across it. Since the anode of V_2 is negative with respect to its cathode it does not conduct. During the following negative half cycle 2, the voltages are reversed and V_2 conducts while V_1 does not. Therefore during a negative half cycle current flows through the other half of the secondary winding BC and through R in the same direction as during the positive half cycle and consequently another pulse of positive voltage appears across R.

The complete description of the action of the circuit is summarized graphically in Fig. 3-3 (c). The first graph shows the complete a-c secondary voltage from A to B. The second and third graphs show the applied voltage across V_1 and V_2 respectively using the point C as reference point. The fourth graph shows that the output voltage across the load resistor R consists of a series of positive pulses at twice the frequency of the input a-c voltage. It should also be noticed that the amplitude of this voltage is one-half the amplitude of the total secondary voltage. This occurs because only one-half of the secondary winding is used to provide the applied voltage when either valve is conducting. The full-wave rectifier is at a slight disadvantage with the half-wave rectifier in this respect because it requires a transformer which has twice the secondary voltage as that required for a half-wave rectifier for a given rectifier output. For example, suppose a peak voltage of $500V$ is required in the output of a rectifier. Neglecting voltage drop in the rectifier valve, a half-wave rectifier needs a transformer which delivers $500V$ peak, but a full-wave rectifier needs a transformer with a $1000V$ peak secondary output since only $500V$ is applied across each half of the full-wave rectifier circuit. This disadvantage is greatly offset by the fact that the full-wave rectifier supplies a more continuous current which is more easily smoothed out than the cur-

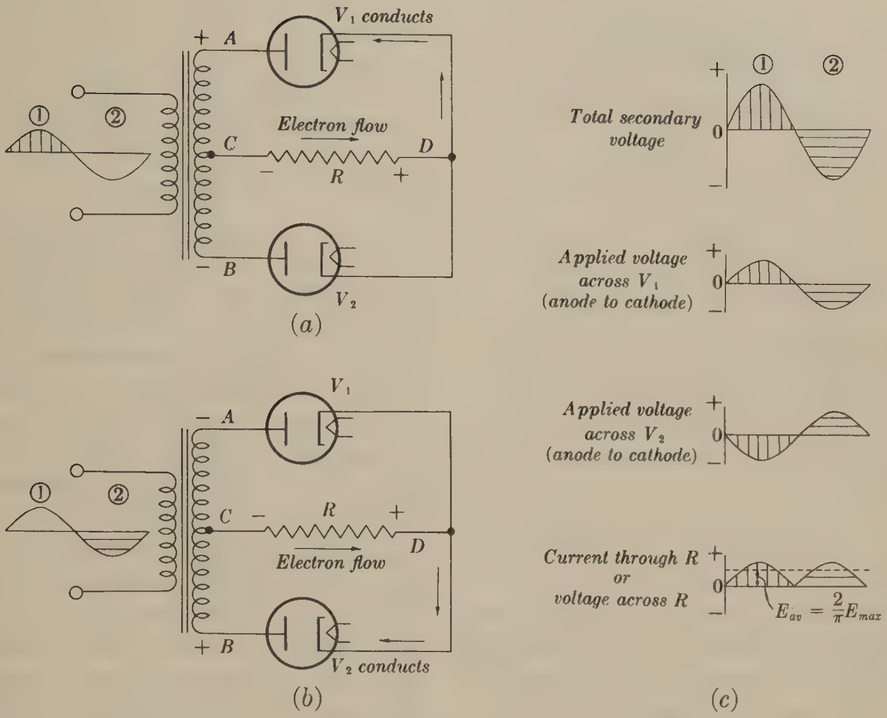

V_1 conducts
+ A
①
②
Electron flow
C
D
−
R
+
− B
V_2
(a)
V_1
− A
①
②
C −
R
+
D
Electron flow
+ B
V_2 conducts
(b)
Total secondary voltage
Applied voltage across V_1 (anode to cathode)
Applied voltage across V_2 (anode to cathode)
Current through R or voltage across R
+
0
−
$E_{av} = \frac{2}{\pi}E_{max}$
(c)

FIG.3-3 A simple full-wave rectifier circuit with voltage and current waveforms.

rent pulses from the half-wave rectifier. Full-wave rectifiers are used almost exclusively in the H.T. power supplies of all amplifiers, radio receivers and transmitters.

The circuit diagram of a simple full-wave rectifier as shown in Fig. 3-3 (a) is best for explaining the action of the circuit, but in actual practice the two diodes are usually included in one valve, a double diode, and the circuit appears as in Fig. 3-4. The points labelled, A, B, C and D correspond in both Figs. 3-3 (a) and 3-4 so that the identity between the two diagrams may be seen. The winding L_1 is used solely for heating the cathode of the valve. L_2 is the high-voltage winding which supplies the a-c current to be rectified. The load resistor R may be connected to the middle of L_1 to prevent any additional variations in voltage which would be produced by the voltage drop across the cathode in a directly-heated valve. However, this is usually not worth the trouble and it is ordinarily connected to one side of the cathode.

3.3 Filter circuits. The d-c outputs of the half-wave and full-wave rectifiers described in the previous section may be used directly for applications such as charging batteries or providing a direct current for relays in control circuits. However, the d-c outputs are far from

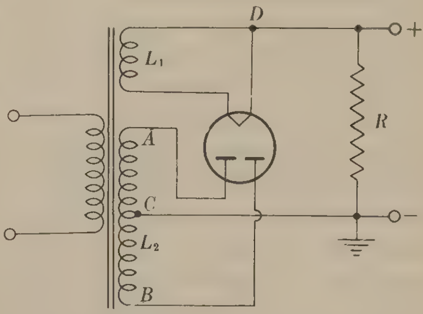


FIG.3-4 A conventional full-wave rectifier circuit.

being smooth d-c, and consist of pulses of current which produce positive pulses of voltage across the load resistor. The voltage has a value equal to $\frac{1}{\pi}$ of the peak value for half-wave rectification and $\frac{2}{\pi}$ of the peak voltage for full-wave rectification. The variation in voltage above and below the average value is called *ripple.* In all radio equipment the H.T. power supply is used to apply a steady d-c voltage to the anodes of the valves and only a very small amount of ripple is permissible. Consequently the output of a rectifier must be smoothed out so that a steady d-c voltage, with very little ripple present, is applied to the valves. This is accomplished by means of *filter circuits.* Filter circuits in general consist of a capacitor or capacitors combined with inductors or resistors. The various types of circuits used to perform this filtering action are discussed in the following sections.

A. The Simple Capacitor Filter

This filter consists of a single capacitor which is connected in parallel with the load (Fig. 3-5). Briefly, the capacitor reduces the fluctuations in the output voltage because it is charged and therefore stores up energy while the rectifier is producing its pulse, and gives up the energy again by discharging through the load between the pulses. In particular, consider what happens when the output of a half-wave rectifier is applied across this filter circuit. During the first positive pulse, as the voltage rises, the capacitor is charged up until the peak voltage is reached. The anode voltage then falls sinusoidally but the cathode potential remains about the same due to the charge held on the capacitor. Thus the anode potential soon falls below the cathode potential and the valve stops conducting shortly after the peak. The capacitor can then only discharge through the load resistor R. The rate of discharge depends on the RC time constant of the circuit. A large capacitance is chosen so that this time constant is long in comparison with the period of the input a-c voltage. Thus the capacitor continues to discharge slowly through R and the voltage across the load falls very little during the negative half cycle. On the next pulse, when the anode voltage of the diode rises above the cathode potential, the valve again conducts and recharges the capacitor practically to the peak of the applied voltage. Then a new cycle begins and the process is repeated. The result is that an almost steady d-c voltage, practically equal to the peak of the applied voltage, is maintained across the load.

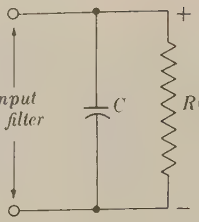


FIG.3-5 A simple capacitor filter circuit.

If the output of a full-wave rectifier is applied to this filter the fluctuations will be much more effectively smoothed out than for the half-wave rectifier. This is so because there is much less time in which the capacitor can discharge between the pulses. Consequently the voltage does not fall as much before a new pulse occurs and recharges the capacitor. As a result the d-c voltage across the load has less ripple remaining in it and also approaches the peak value of the applied voltage more closely than is the case with the output from a half-wave rectifier. This consideration indicates that the effectiveness of the filter depends on the frequency of the input voltage. The higher the frequency the less time there is between pulses and the output from the filter has correspondingly less ripple. In addition a smaller capacitance can be used with higher frequencies since the time constant of the filter circuit does not need to be as long. For this reason the input a-c to many sets designed for the Naval Service is 500 c/s rather than the usual 60 or 25 c/s encountered on shore.

The action of the capacitor filter is shown graphically in Fig. 3-6 (a) and (b), the waveforms for the half-wave rectifier and the full-wave rectifier being shown in (a) and (b) parts respectively. The dotted waveform represents the input voltage to the filter from the rectifier and the solid curve represents the voltage across the capacitor and load. At point *A* the capacitor has been charged to the peak of the applied voltage. The anode voltage then falls rapidly as shown by the dotted waveform and the valve stops conducting. The capacitor then discharges through *R* and the voltage falls slowly until point *B* is reached. At this instant, time t_1 on the graph, the rising anode voltage is equal to the capacitor voltage, i.e. the cathode potential, and the valve begins to conduct. Current continues to flow through the valve until time t_2, point *C* on the graph, when the anode voltage again falls below the cathode voltage and the capacitor can only discharge through *R*. The process is again repeated at points *D* and *E* and the output from the filter is seen to be a d-c in which the fluctuations have been very much reduced. It should also be noticed that since the time from *A* to *B* and from *B* to *C* are both shorter in graph (b) than graph (a) that the voltage output from the filter having a full-wave rectifier input is smoother and slightly higher than that having a half-wave rectifier input.

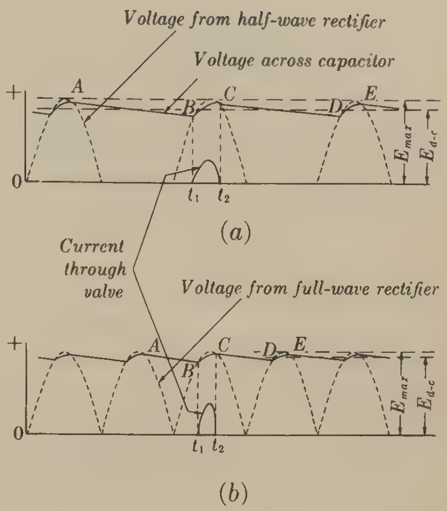


FIG. 3-6 Capacitor-filter voltage and current waveforms.

One of the characteristics of this type of filter is that it causes the rectifier valve to conduct heavily for only a short time during each cycle. During this time the valve must carry sufficient current to maintain the current through *R* and also recharge the capacitor to the peak of the applied voltage. Thus the peak current through the valve is very much larger than the average load current through *R*. Care must be taken to ensure that this peak current does not exceed the peak current rating of the valve. For a given load resistor the size of the peak current is controlled by the size of *C*. If *C* is large the capacitor voltage falls very little between the pulses because of the long time constant, but if *C* is small the voltage falls considerably. For this reason the large capacitor loses more charge than the small capacitor during the same period. Fig. 3-7 aids in the explanation of this fact.

In Fig. 3-7 (a) the dotted curve represents the output voltage of a rectifier across a pure resistance load, i.e. with no filter. The solid curve *A* represents the voltage across *R* when a large filter capacitor is used and curve *B* represents that for a smaller capacitor. The valve conducts only during the time that the

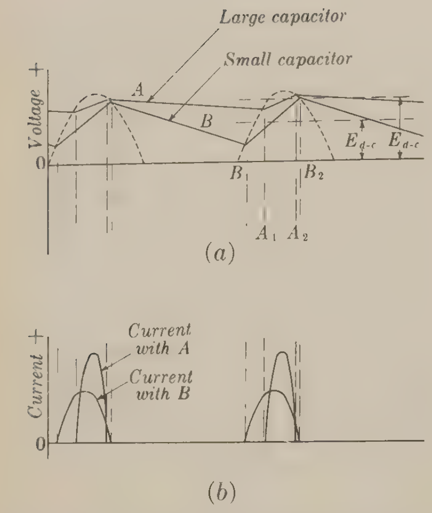


Fig.3-7 The effect of the size of the filter capacitor on the output of a capacitor filter.

applied voltage is greater than the voltage of the capacitor. This occurs between times A_1 and A_2 in Fig. 3-7 for a large capacitance and between times B_1 and B_2 for the smaller capacitance. The time A_1A_2 is shorter than the time B_1B_2. Thus for the large capacitance the valve must replace more charge in a shorter space of time than for the small capacitance and therefore the peak current is higher, Fig. 3-7 (b). The practical result of this is that in choosing the capacitance of a capacitor for a filter a compromise must be made between the amount of ripple in the output and the peak current through the valve.

After the capacitor is chosen for a filter the percentage ripple depends only on the size of the load resistance. If this resistance becomes less, more current is drawn from the supply and the time constant of the capacitor and load becomes smaller. Consequently the capacitor discharges more completely during each cycle as shown in Fig. 3-7 (a). Thus in a capacitor filter the amount of ripple increases with increased load current drawn from the supply. In addition, since the effective d-c voltage is mid-way between the upper and lower variations of the ripple voltage, the output voltage of the filter is lower when the ripple is greater, Fig. 3-7 (a). Thus drawing a greater load current not only increases the ripple but also lowers the filter output voltage across the load.

Besides the above physical explanation of capacitor filtering it is interesting and valuable to understand the action of this filter from a mathematical viewpoint. A mathematical analysis of the pulsating direct current output of a rectifier shows that it may be considered to be made up of a steady d-c component and an a-c component as shown in Fig. 3-8. The a-c component is not a pure sine wave. An a-c which is not a pure sine wave consists of a sine wave at a certain fundamental frequency combined with sine waves at higher frequencies which are multiples of the fundamental and are called *harmonics.* In the ouput of a rectifier, the fundamental a-c frequency is called the *ripple frequency.* In a half-wave rectifier the ripple frequency is the same as that of the input a-c and in a full-wave rectifier it is twice the input frequency. When the output of the rectifier consisting of a direct current and many a-c components is applied across the capacitor and resistor in parallel the

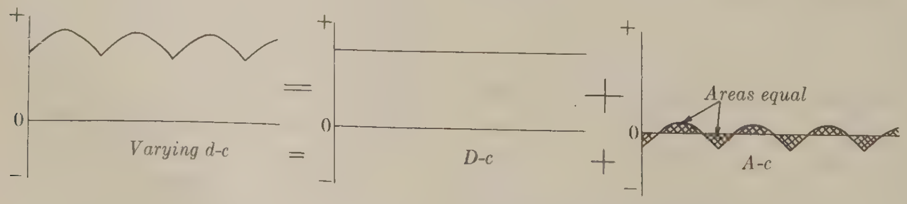


Fig.3-8 The components of a varying d-c voltage.

d-c can only flow through R but the a-c can flow through either C or R. The capacitance of C is chosen so that it provides a path for the a-c of low impedance compared to R. Consequently very little a-c voltage is developed across R and thus the capacitor has prevented the ripple from appearing across R or in other words it has *filtered* the a-c from the output.

One of the advantages of this mathematical approach is that the amount of ripple remaining in the output may be stated exactly. It is usually expressed as a percentage and is equal to

$$\frac{\textit{r.m.s. value of the ripple voltage}}{\textit{output d-c voltage}} \times 100$$

This is a measure of the effectiveness of the filter. The percentage ripple in a power supply used with c-w transmitters may be as high as 5%. For receivers and transmitters used for r-t the percentage ripple must be of the order of .25%. A higher percentage ripple than this results in a steady hum at the ripple frequency in loudspeakers and earphones.

It is possible to calculate the relative a-c impedances of the load resistor and the filter capacitor and therefore obtain an estimate of the amount by which the ripple is reduced by a filter. For example, in a full-wave rectifier the fundamental ripple frequency is 120 c/s, $C = 16\mu F$, $R = 5000\Omega$.

Impedance of C is

$$X_c = \frac{1}{2\pi fc}$$

$$\therefore\ X_c = \frac{10^6}{2\times3.14\times120\times16}$$

$$\doteqdot 90$$

$\therefore$ impedance of C is approximately 90Ω. Therefore the a-c impedance of the output with no filter is 5000Ω and the a-c impedance of the output with the filter is very nearly 90Ω. Therefore the a-c voltage developed across the load with the filter present is $\frac{90}{5000}$ or approximately $\frac{1}{50}$ of the a-c voltage developed with no filter present.

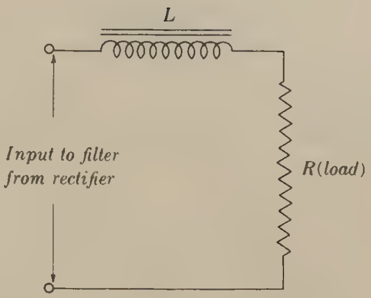


FIG.3-9 A simple inductor filter circuit.

Thus the addition of the filter reduces the ripple voltage to $\frac{1}{50}$ of its former value.

B. *The Simple Inductor Filter*

The simplest inductor filter consists of a single inductor connected in series with the load resistor as in Fig. 3-9. Since an inductor opposes any change in the magnitude of the current flowing through it, it reduces the fluctuations in the pulsating current produced by the rectifier. This action may be considered in slightly more detail as follows. When the current through the inductor increases, a magnetic field is built up around it. When this occurs, two things happen. First a back e.m.f. is produced which prevents the current from rising as much as it normally would; and secondly, energy is stored up in the magnetic field. After the rectifier current reaches its peak and starts to decrease, the magnetic field falls, thereby supplying energy to the circuit. The forward e.m.f. produced by the falling magnetic field keeps the current from falling as rapidly as it would if no inductor was in the circuit. Thus it may be seen that the current cannot either rise or fall as rapidly when the inductor is in series with the load, and thus the ripple in the output is reduced. The larger the inductance used in the filter the better will be the smooth-

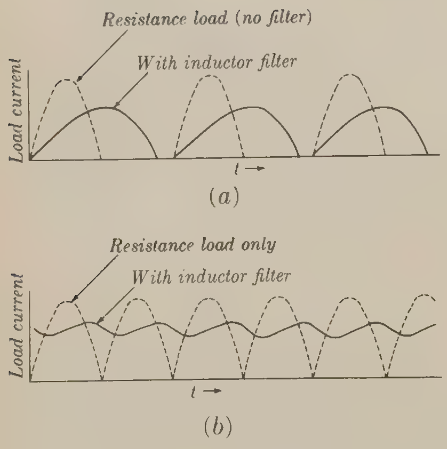


Fig. 3-10 Inductor-filter voltage and current waveforms.

ing action. However, there is a practical limit to the size of inductance which can be used since the d-c resistance of the inductor winding is excessive if it has very large inductance and the IR drop in this resistance lowers the output voltage across the load.

The graphs in Fig. 3-10 (a) and (b) show the effect produced on the load current by the inductor filter when it is used with a half-wave rectifier and a full-wave rectifier respectively. The dotted curves represent either the voltage input to the filter from the rectifier or the load current with a pure resistance load. The solid curves represent the load current when the inductor filter is present. Since the current never rises to the peak value that is reached without the inductance, the output voltage also does not reach the peak of the applied a-c. Consequently the output voltage from a rectifier is lower with an inductor filter than with a capacitor filter. However the inductor filter has several advantages against this disadvantage. Increasing the load current has little effect on the amount of ripple in the output and therefore an increase in load has much less effect on the output voltage in an inductor filter than in a capacitor filter. In addition there is no problem of a peak current through the valve since the current through the valve also flows through the inductor and the load. For these reasons an inductor filter is used wherever a large current drain is needed from the power supply.

As in the case of the capacitor filter, the action of the inductor filter may be considered from a mathematical viewpoint. The action depends on the relative impedances of the inductor and resistor to the d-c and a-c components of the rectifier output. As far as the d-c is concerned, the inductor offers very little resistance and most of the d-c voltage appears across the load resistor. However, in the case of the a-c the inductor offers a higher impedance than does the resistor and consequently more a-c voltage is developed across it than across the load. Since the inductor stops or chokes the ripple voltage from appearing across the load, inductors which are used in filter circuits are commonly called *chokes*. A numerical example will show the effectiveness of this type of filter. Suppose a 20 H choke is connected in series with a 5000Ω load in the output of a full-wave rectifier whose input frequency is 60 c/s.

$$\text{Ripple frequency} = 120 \text{ c/s}$$

$$\begin{aligned} \text{Also} \qquad X_L &= 2\pi f L \\ &= 2\times 3.14\times 120\times 20 \\ &\doteqdot 15{,}000 \end{aligned}$$

$\therefore$ impedance of the choke is 15,000Ω.

Therefore for a given ripple voltage $\frac{3}{4}$ of the a-c voltage appears across L and $\frac{1}{4}$ of the a-c voltage appears across R.
Therefore the ripple has been reduced to $\frac{1}{4}$ of its initial value in the output.
This shows that a very large inductance is necessary if good filtering is to be obtained.

C. *Inductor-Capacitor Filters*

If good filtering is required, the use of a capacitor or an inductor alone in a filter circuit is not practical because of the large capacitance or inductance which is needed. The disadvantages resulting from using large values of capacitance or inductance are discussed in the previous sections. To provide good filtering with practical sizes of these components various combinations of inductors and capacitors are used.

1. CHOKE-INPUT FILTER.

The simplest combination filter consists of an inductor in series with the load and a capacitor in parallel with the load. This is known as a *choke-input* or *inductor-input* filter (Fig. 3-11) since the inductor is the first element of the filter following after the rectifier. The inductor prevents changes in current through the load and the capacitor smooths out voltage variations across the load. The combined action of these components eliminates the ripple to a great extent.

The amount of reduction in ripple voltage by this type of filter may be estimated by the impedances in the circuit. For example, if the input to the filter of Fig. 3-11 has a ripple frequency of 120 c/s, $L = 10\,H$ and $C = 8\mu F$

then $X_L = 2\pi f L$

$= 2 \times 3.14 \times 120 \times 10$

$\doteqdot 7540$

and $X_C = \dfrac{1}{2\pi f C}$

$= \dfrac{1}{6.28 \times 120 \times 8 \times 10^{-6}}$

$= \dfrac{10^5}{6.28 \times 12 \times 8}$

$\doteqdot 166.$

Therefore the inductive reactance is 7540 ohms and the capacitive reactance is 166 ohms at the ripple frequency. The reactance of L is about fifty times that of C. Therefore about 98% of the a-c voltage in the circuit appears across the inductor and only about 2% remains across the capacitor and load. Thus the ripple voltage in the output is reduced to $\frac{1}{50}$ of its value in the input to the filter, with much smaller values of inductance and capacitance than were used in the simple inductor or capacitor filter.

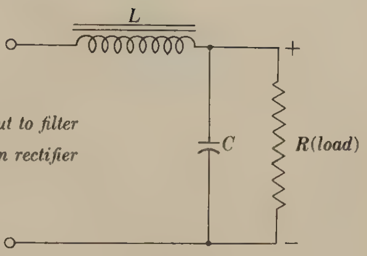


FIG.3-11 Choke-input filter circuit.

In choke-input filters the inductor-filter properties predominate. Thus the voltage output is much lower than the peak value of the applied a-c, the peak current through the valve is not excessive and the percentage ripple in the output is nearly constant with increasing load. If very good filtering is necessary another inductor and capacitor may be added as in Fig. 3-12. L_1 and C_1 form the first section and L_2 and C_2 the second section of the filter. Because of its shape, each section is called an *inverted-L section*. The whole filter is thus called a *double inverted-L section filter*. The action of the second section of the filter is similar to the action of the first section. A large percentage of the ripple voltage left across C_1 is developed across L_2.

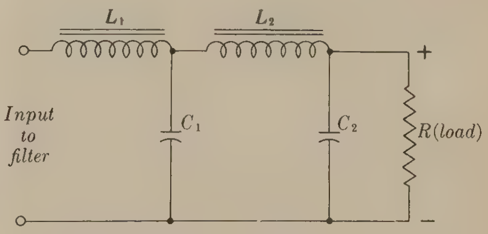


FIG.3-12 Double inverted-L section filter.

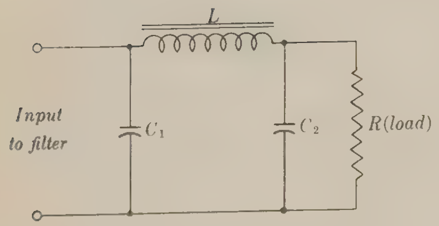


FIG.3-13 Capacitor-input filter circuit.

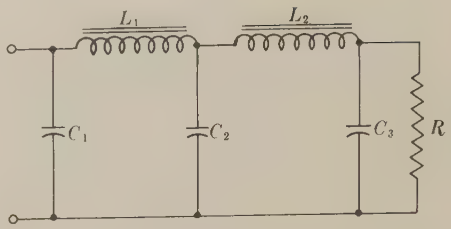


FIG.3-14 π-section filter with additional inverted-L section.

This leaves only a small fraction of the ripple voltage across C_1 to appear in the output voltage across C_2. Suppose that each section reduces the ripple voltage to 2% of its value in the input. Thus only 2% of the original ripple voltage appears across C_1 and only .04% of the original ripple voltage appears across C_2. If necessary, filtering may be made more complete by adding more sections. However each additional section causes a drop in the output voltage because of the IR drop across the inductor.

2. CAPACITOR-INPUT FILTER.

A filter having a capacitor as its first component is called a *capacitor-input filter.* Fig. 3-13 shows a commonly used combination capacitor-input filter. It consists of an input shunt capacitor, a series inductor and a capacitor across the load. From its shape it is called a *π-section filter.*

The action of the circuit may be understood by considering the impedances presented to the a-c and d-c components of the input by the various branches. The d-c can only flow through L and R. The resistance of the windings of L is very low so that practically all of the d-c voltage appears across R. For the a-c however, the inductance offers a high-impedance and the capacitance a low-impedance path. Consequently a large proportion of the a-c is shunted by the capacitor C_1. The small amount of a-c which flows through the inductance is again confronted by a high-impedance path through the load and a parallel low-impedance path through C_2. Consequently very little a-c flows through R and there is very little ripple in the output. The input capacitor may be smaller in the π-section filter than in the single-capacitor filter and this reduces the danger of excessive peak currents through the valve. If additional filtering is necessary an inverted-L section may be added to the circuit as in Fig. 3-14. As before, the output voltage is reduced by the IR drop in the additional choke. Consequently this addition is made only when absolutely necessary.

In a capacitor-input filter, the properties discussed under capacitor filtering predominate. The output voltage approaches the peak value of the applied a-c, the peak current through the valve is high during the short conducting period and the ripple voltage increases with increased load current.

D. Resistor-Capacitor Filter

In capacitor-input filters, a resistor is sometimes used instead of an inductor (Fig. 3-15). The action of such a circuit is exactly the same as that of the capacitor-input filter previously described. The main advantage resulting from the use of a resistor is the saving in money and space. However the use of this type of filter is limited by two considerations. First, the resistor is not as effective as the inductor in reducing the ripple. Secondly, the circuit may not be used when even a medium load current is required because the large IR drop

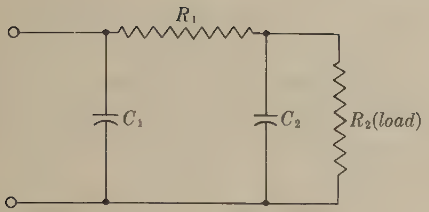


FIG.3-15 π-section resistor-capacitor filter.

across R_1 lowers the output voltage across R_2.

The resistor-capacitor filter is used in the power supplies of oscilloscopes and cheaper makes of commercial a-c d-c receivers where a small amount of ripple is permissible and the load current is small. Aside from its application in power supplies it is often used as a *decoupling network* to prevent feedback from one stage of an amplifier to the previous ones through the anode circuit. It is also employed as a filter to keep the screen grid of a pentode at a constant potential.

E. Voltage Regulation

When the load current drawn from a power supply is increased, the output voltage usually falls. The constancy of this d-c voltage for varying values of load current is called the *regulation* of the power supply. Thus a power supply which undergoes a large drop in output voltage when the load current is increased is said to have *poor voltage regulation.* This fall in voltage is caused partly by the increased IR drop in the transformer winding, the anode resistance of the valve and in the choke coil of the filter and partly by the characteristics of the particular type of filter circuit being used.

For the purpose of comparing different power supplies, the drop in voltage for a certain specified change in load current is indicated by a number called the *percentage regulation.* This figure is obtained by expressing the change in voltage as a percentage of the initial voltage. Percentage regulation is equal to

$$\frac{E_1 - E_2}{E_1} \times 100$$

where

E_1 = the no-load voltage in volts

E_2 = the full-load voltage in volts.

The graphs in Fig. 3-16 show how the output voltage varies for different load currents when capacitor-input and inductor-input filters are used with a full-wave rectifier.

The capacitor-input filter produces a d-c voltage output at no load which is practically equal to the peak value of the input a-c. As the load current is increased however the d-c output voltage steadily decreases. This is due not only to the IR drops in the circuit but also to the fact that the capacitor must discharge more completely to supply a larger load current. Thus, a capacitor-input filter has poor regulation. This type of filter may be used for small loads such as are found in receivers or for a constant load. The decrease in output voltage with increasing load current may be reduced by increasing the capacitance. However this may result in an excessive peak current through the valve during the charging of the larger capacitor.

The inductor-input filter also produces under no-load conditions a d-c voltage

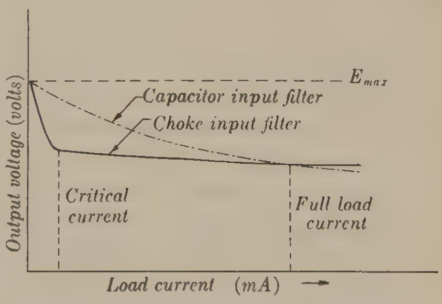


FIG.3-16 Regulation curves for capacitor-input and choke-input filters.

output which is nearly equal to the peak value of the a-c input. However, as the load current increases, the output voltage drops quickly to a point beyond which the output voltage levels off. The current drawn by the load at this point is called *the critical current.* For currents greater than the critical current further slight decreases in output voltage are caused only by the increased *IR* drops in the circuit. Thus good regulation is obtained for loads greater than the critical current. For a given load, the larger the inductance the more quickly the critical current is reached. In practice a *load* resistor is placed across the output terminals of the choke-input filter. This resistor is the correct size to cause the critical current to flow in it at all times. This prevents the output voltage from rising, if the external load current is small, and putting too great a voltage on the filter capacitors. With this method fairly good regulation is obtained for varying values of external load current.

If good regulation is required over a wide range of load current, a *swinging choke* is very useful. Such a choke has high inductance with small d-c load currents and a much lower inductance with large direct currents due to partial saturation of the core. This property is valuable in producing good regulation for the following reasons. In a choke-input filter the larger the inductance, the smaller is the critical current. However, with a small inductance the choke-input filter partially assumes the properties of a capacitor-input filter and the output voltage rises. Thus the swinging choke provides good regulation because with small load currents the inductance is high, the critical current is very small and therefore the output voltage is prevented from rising rapidly. With large load currents the inductance decreases, the filter begins to assume the properties of a capacitor-input filter and the output voltage rises sufficiently to counteract the fall in voltage caused by the *IR* drops in the circuit. Thus a swinging choke keeps the output voltage quite constant over a wide range of load currents.

The construction of a swinging choke is similar to an ordinary a-f choke except that the core has a very small air gap or no air gap at all. This fact causes the inductance to be greater for the same physical size as an ordinary choke when little d-c current is drawn but causes the core to saturate more easily and therefore lower the inductance when larger d-c currents are drawn by the external load.

F. Location of the Inductor

A series inductor such as is used in the filters described above may be placed either in the grounded lead or in the lead which is off ground. Two factors, arising from the fact that the iron core of this inductor is usually grounded, must be considered. They are:

(i) the distributed capacitance from turns to ground

(ii) the insulation between the turns and ground.

Suppose the inductor is placed in the lead which is off ground. In Fig. 3-17 (a) it may be seen that the distributed capacitance is shunted across the line and assists in the filter action. On the other hand the insulation between the turns and the core must be strong enough to prevent breakdown due to the high

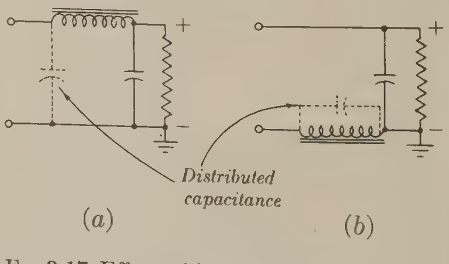


Fig.3-17 Effect of location of inductor in the H.T. or ground line.

voltage of the power supply. In a large supply this may prove expensive.

Suppose the inductor is placed in the grounded lead, Fig. 3-17 (b). In this case the distributed capacitance shunts across a portion of the filter inductor and acts to hinder the filter action since it reduces the impedance of the inductor. However, there is little voltage drop between the coils and ground and little danger of breakdown of insulation even in a high-voltage power supply.

G. Resonance Effect in a Filter Circuit

This effect may occur in a choke-input filter where the first inductor and the first capacitor are in series across the output of the rectifier circuit. If they are in resonance at the ripple frequency an acceptor circuit is formed which passes a very high current. This current causes a large voltage at the ripple frequency to appear across the capacitor. Thus the ripple is increased and the effect is opposite to that desired in filtering. The rectifier valves may be damaged either by the large peak current or by the large inverse voltage produced. Since the resonant frequency in a series circuit depends on the product LC, the inductance and capacitance must be chosen to have a product of at least twice the value for resonance at the ripple frequency to prevent resonance effects.

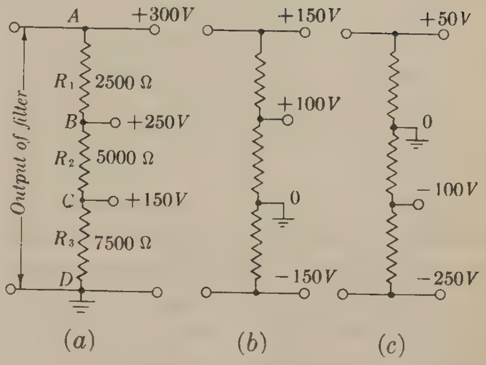


FIG. 3-18 (a) Simple voltage divider (b) and (c) Effect of changing position of ground.

3.4 Voltage dividers. Most power supplies have a resistor connected across the output terminals in parallel with the load. There are three main reasons why this is done and the resistor is named according to the principal reason for which it is being used. If it serves the purpose of discharging the filter capacitors after the rectifier has been turned off it is called a *bleeder resistor*. If it is used at the output of a choke-input filter to draw the critical current and therefore improve the voltage regulation it is called a *load resistor*. If the resistor is tapped at several points to provide a variety of voltages which are less than the total output voltage it is then called a *voltage divider*. In most cases a resistor placed across the output terminals of a power supply fulfils more than one of these functions. However, if it serves as a bleeder resistor only, it usually has a very high resistance so that it draws very little current and therefore does not waste power. If it is used as a load resistor, it is designed to draw about 10% of the full load current and the proper wattage rating must be allowed for. A load resistor may also be used very conveniently as a voltage divider. A simple voltage divider is shown in Fig. 3-18. An example of the method of calculating the size of resistors needed to provide certain voltage outputs is given in the following paragraphs.

Suppose a rectifier power supply is designed to deliver $200 mA$ to a load at a voltage of $300 V$ and in addition voltages of $250 V$ and $150 V$ are to be tapped from a voltage divider as shown in Fig. 3-18 (a). The voltage divider is divided into three sections to provide the other two voltages. The resistance of the whole voltage divider is first found and then it is divided into the parts R_1, R_2, R_3 to provide the correct voltages. The voltage divider is chosen so that about 10% of the total load current flows through it.

Thus for an output of $200mA$ the current through it is about $20mA$ and the total resistance is

$$R = \frac{E}{I} = \frac{300}{20\times10^{-3}}$$

or 15,000 ohms. Since the voltage drop from A to B is $50V$ or $\frac{1}{6}$ of the total voltage, then R_1 is $\frac{1}{6}$ of 15,000 or 2,500 ohms. Another drop in voltage of $100V$ or $\frac{1}{3}$ of the rectifier output voltage appears across R_2, therefore R_2 is $\frac{1}{3}$ of 15,000 or 5,000 ohms. Since R_3 has a voltage drop across it of $150V$, which is $\frac{1}{2}$ of the output, its value is 7500 ohms. These output voltages of $250V$ and $150V$ are obtained only if no current is drawn from the taps at B and C but they are independent of the load current drawn at A provided the power supply has good regulation. This method is used to supply bias voltages in circuits where no current is drawn from the source. In this connection, the position of ground may be taken at any point on the voltage divider provided there is no other ground in the circuit. In this way various potentials above and below ground are obtained as shown in Fig. 3-18 (b) and (c).

If it is desired to supply $250V$ and $150V$ to loads which draw current, then R_1, R_2 and R_3 must be recalculated, since with the previous sizes of R_1, R_2 and R_3 the voltages at B and C change when current is drawn. If a load L_1 draws a current of $50mA$ from the $250V$ tap and a load L_2 draws a current of $10mA$ from the $150V$ tap (Fig. 3-19) and the bleeder current with full load is $20mA$ then the new values of R_1, R_2 and R_3 are calculated as follows. If the bleeder current is $20mA$ then the current in R_3 is $20mA$, therefore $R_3 = \frac{150}{20}\times1000 = 7500\Omega$. Since $10mA$ is drawn from point C therefore the total current through R_2 is $30mA$ and $R_2 = \frac{100}{30}\times1000 \div 3333$ ohms. Similarly the total current through R_1 is $80mA$ and $R_1 = \frac{50}{80}\times1000 = 625\Omega$.

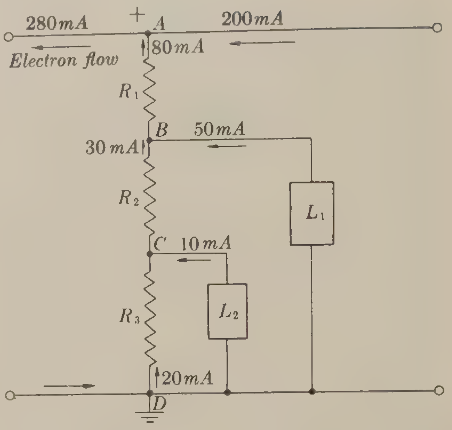


FIG.3-19 Voltage divider showing the effect of loads on the voltage division.

The values of R_1, R_2 and R_3 are then 625, 3333 and 7500 ohms respectively to give output voltages of $250V$ and $150V$ at the two taps for the specified currents.

In choosing a resistor for a voltage divider, the power to be dissipated in the resistor must be known. In the above case, R_1 is 625 ohms and has $80mA$ flowing in it. The power dissipated is then

$$P = I^2R = (80\times10^{-3})^2\times625 = 4W.$$

To provide a margin of safety, a resistor capable of dissipating at least double the power is always chosen, consequently R_1 is a 10-watt resistor. Similarly R_2 and R_3 each dissipate 3 watts under normal load conditions and therefore they also are 10-watt resistors.

3.5 A receiver power supply. In Fig. 3-20 a receiver power supply for use with smaller receivers is shown. The power transformer is supplied from the a-c mains through the on-off switch S_1. There are three secondary windings in this transformer and in some cases there may be more. The centre winding is the H.T. winding, centre-tapped for full-wave rectification. The other two windings are filament windings, one for the rectifier valve and one for the valves in the receiver. The rectifier valve usually

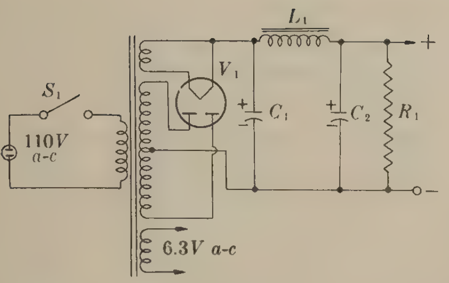


FIG. 3-20 A conventional receiver power supply.

requires a separate winding for its filament because of the large current drain of about 2A. Since the other receiver valves only require about 0.3A they can be supplied by one winding.

The capacitance of the input capacitor C_1 in the π-section filter is usually about 4μF as compared to 8μF for the output capacitor C_2. The input capacitance is frequently smaller to avoid peak currents that are too high for the valve. The size of the inductance is usually between 20 and 30H. The field coil of a dynamic speaker is often made to act as this inductor. In that way two requirements are filled at the same time; the current from the rectifier is filtered by the inductance of the coil and also this current produces the magnetic field for the speaker. The above sizes of filter components are typical for a ripple frequency of 120 c/s. However, for any other ripple frequency the size may be multiplied by the factor $120/f$ where f is the frequency.

In this circuit a bleeder resistance R_1 is connected across the output of the power supply. Its function in this circuit is to allow C_1 and C_2 to discharge when the supply is turned off. If R_1 is not in the circuit, C_1 and C_2 retain their charges for a considerable time and are dangerous to handle even if the switches are off.

Typical outputs of receiver power supplies with π-section filters are as follows. With a centre-tapped transformer secondary voltage of 425-0-425V r.m.s., the voltage output is usually about 400V d-c with a current drain of about 170mA. When the transformer voltage is 300V r.m.s. on each side of the centre tap, the output voltage obtained is about 275V with a current drain of 120mA.

The circuit of Fig. 3-21 shows a power supply suitable for use with larger receivers requiring up to 200mA at 325V. The voltage and circuit component values are shown throughout the circuit. There are several additional features in this power supply which possibly need explaining.

(i) First, this circuit shows an example

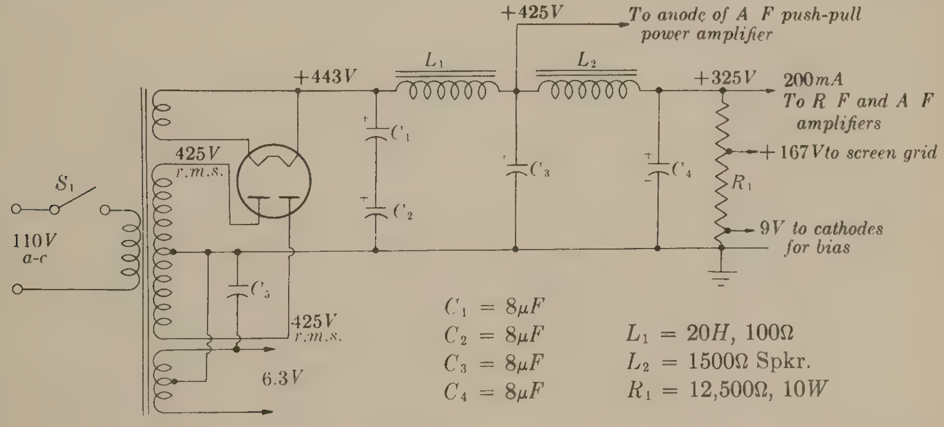


FIG. 3-21 Power supply suitable for larger receivers.

of *graded filtering* in which the H.T. to the amplifier stages with low power level is better filtered than the H.T. to the later or high-power stages. The earlier stages of amplification need a very smooth H.T. since any hum is greatly amplified by each succeeding stage. In this circuit they are supplied by the 325 V output in which the hum has been reduced by an additional filter stage. However, the 425 V taken after the first stage of filtering is smooth enough for use with a push-pull amplifier output stage. Taking the 425 V output after L_1 saves power because less current flows through the higher resistance choke L_2. In addition, a higher voltage is obtained for the power stage while still using the same power supply.

(ii) Second, the reason for using C_1 and C_2, two $8\mu F$ capacitors in series rather than one $4\mu F$ capacitor is that the capacitor in this position must be able to withstand a peak voltage of approximately 600 V. Since standard capacitors are manufactured to withstand only 450 $W.V.$, it is cheaper to use two standard $8\mu F$ capacitors in series giving $4\mu F$ which will withstand 900 V rather than to install a specially constructed $4\mu F$ capacitor to withstand this voltage.

(iii) Third, the capacitor C_5 is needed to by-pass any r-f, which may get into the filament circuit, to ground.

3.6 Bridge rectifiers. In the full-wave rectifier described in section 3.2 C it is necessary to have a transformer with a secondary capable of developing a maximum voltage at least twice as large as the desired d-c voltage output. Transformers capable of a very high voltage output are expensive. The use of an expensive transformer is eliminated by the use of a bridge rectifier because its voltage output approaches the peak voltage of the entire transformer secondary. In addition, the bridge rectifier makes use of the full voltage cycle and therefore it combines the advantages of the full-wave and the half-wave rectifier.

A circuit of this type of rectifier is shown in Fig. 3-22. The action of the circuit may be described as follows. During the half cycle when the upper end A of the transformer secondary is positive with respect to the lower end B, V_1 and V_4 conduct while V_2 and V_3 are cut off. Consequently electrons flow from B, through V_4, the load resistor and V_1 to A and back to B through the transformer secondary. The path is indicated by the solid arrows on the diagram. During the other half cycle, B is positive with respect to A. In this case V_1 and V_4 are cut off and the con-

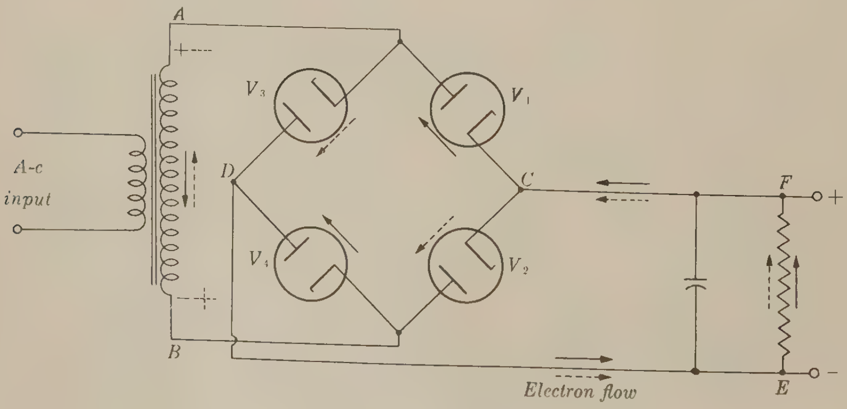


FIG. 3-22 A bridge rectifier circuit.

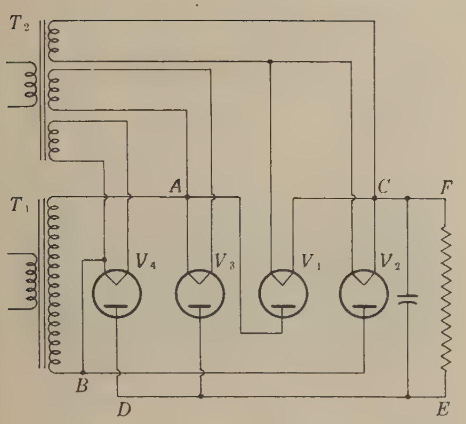


FIG.3-23 Bridge rectifier circuit showing filament connections.

ducting path is through V_3, the load and V_2 as indicated by the dotted arrows on the diagram. In either case the electrons move from E to F through the load and therefore the load resistor voltage always has the same polarity as marked in the diagram. The magnitude of the voltage across the resistor is almost equal to the voltage applied by the whole transformer secondary.

In the diagram it may be seen that the cathodes of only two valves, V_1 and V_2, are at a common potential. Because of this fact, a common filament winding as used in other types of rectifiers cannot be used in this case to heat all four cathodes. Consequently, it is necessary to use three separate windings in the case of directly-heated cathodes and it is desirable to do so with indirectly-heated cathodes. In the latter case, if three separate windings are not used, a large voltage drop exists between the cathode and the heater in two of the valves and there is a possibility of insulation breakdown. Fig. 3-23 shows a wiring diagram for the same bridge circuit having the appropriate filament connections. T_1 is the H.T. transformer and T_2 is the filament transformer.

In a bridge rectifier the ripple frequency, the peak current through each valve and the average current through each valve are the same as for a full-wave rectifier. The only disadvantages of the bridge rectifier are that it is more expensive and occupies more space because two extra valves are needed to produce the same output as a full-wave circuit and extra filament voltage supplies are needed.

One of the main advantages of the bridge rectifier circuit is that it produces the lowest peak inverse voltage across the valves of any of the three main types. The *peak inverse voltage* is the maximum voltage appearing across the valve when it is non-conducting, i.e. when the anode is negative with respect to the cathode. In the simple half-wave rectifier the peak inverse voltage is equal to the peak of the a-c from the transformer secondary. If, however, a capacitor filter is placed across the load, then the inverse voltage is equal to the voltage held on the capacitor plus the voltage across the secondary or very nearly twice the peak of the applied a-c (Fig. 3-24). The peak inverse voltage in a full-wave rectifier is the same as that in a half-wave rectifier for the same voltage output, since each half of the circuit is a half-wave rectifier operating at half the total voltage across the secondary which is twice that in a single half-wave rectifier for the same output voltage. In a bridge rectifier, however, the output voltage is very nearly equal to the total

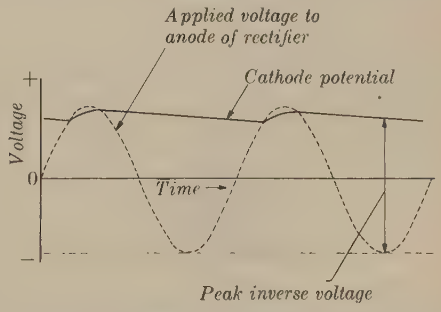


FIG.3-24 Waveforms illustrating peak inverse voltage with a capacitor-input filter.

secondary voltage as in a half-wave rectifier. Thus the peak inverse voltage is equal to the peak secondary voltage or twice this with capacitor filtering. However it is divided equally between the two non-conducting valves in series and therefore the inverse voltage appearing across each valve is only one-half that experienced by a valve in a half-wave or a full-wave rectifier for the same output voltage.

3.7 Metallic rectifiers. Another device which permits current to flow in one direction better than in the other is a *metal rectifier*. The three main types of metal rectifiers are copper oxide, selenium and copper sulphide.

The action of the copper-oxide rectifier depends on the property of copper in contact with a thin coating of cuprous oxide to conduct current much more freely from the metal to the oxide than in the reverse direction. The action of the selenium rectifier depends on the fact that a thin film of selenium on a surface of iron presents a large resistance in one direction and a small one in the other. The copper sulphide rectifier is assembled from disks of copper sulphide and magnesium. The circuit symbol for a metal rectifier is shown in Fig. 3-25. The arrow points opposite to the direction in which electrons flow more freely through the rectifier.

A single element of a copper-oxide rectifier cannot withstand a high voltage. If a high voltage is applied, the thin copper-oxide coating may be broken down. Even if the coating does not break down, excessive currents flow due to the small resistance and a large amount of heat is developed. This heating impairs the efficiency of the rectifier by making the resistance to current flow almost the same in both directions and may even cause chemical changes in the rectifier which result in permanent damage. This means that there is a definite upper limit to the voltage which may be applied to a rectifier of this type. In order that the voltage which may be applied to the rectifier may be raised, a number of copper plates coated with copper-oxide on one side are used in series to form a rectifier unit (Fig. 3-26). These plates are fastened together by a bolt through the centre having around it an insulating sleeve so that the plates are not shorted by the bolt. The plates are separated by means of conducting lead washers. The plates protrude beyond the lead washers like fins in order to permit efficient heat radiation for cooling of the rectifier unit. The soft-lead washer also provides uniform contact between adjacent plates. The current and voltage output of circuits using metallic-oxide rectifiers are limited by the number and area of the plates in the rectifier unit.

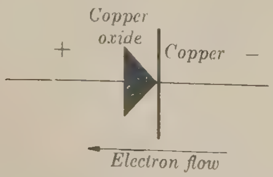


FIG.3-25 Circuit symbol of a metal rectifier.

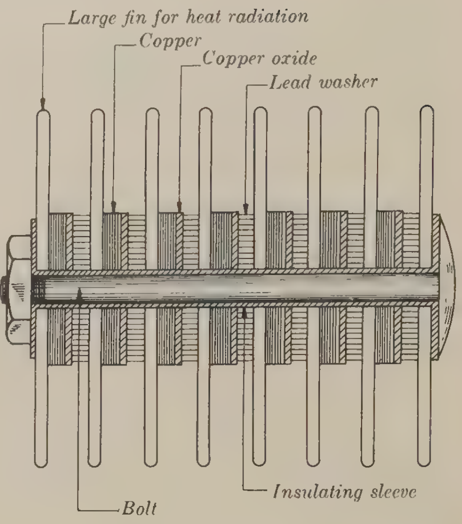


FIG.3-26 Cross-section of a complete metal rectifier unit.

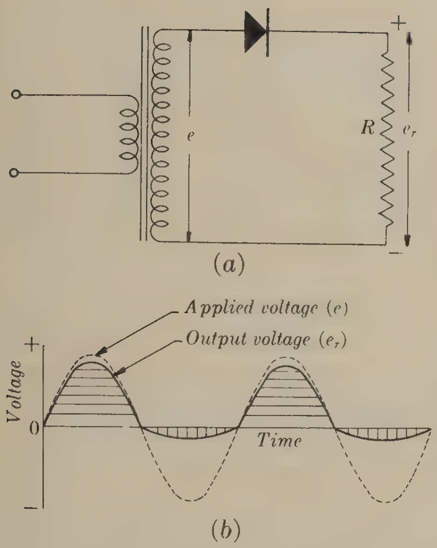


FIG.3-27 Effect of a-c voltage on a metal rectifier circuit.

When an a-c voltage is applied to a metallic-oxide rectifier as in Fig. 3-27 (a) a large amount of current flows in R on the positive half cycle which develops a large voltage of the polarity shown in Fig. 3-27 (b). On the negative half cycle, a small amount of current flows through the rectifier in the opposite direction causing a small voltage across R in the opposite direction, Fig. 3-27 (b). This voltage on the negative half cycle reduces the average value of the output voltage. The amount of reduction depends on the size of the peak inverse voltage of the rectifier and therefore the inverse current. With a capacitor filter in the circuit of Fig. 3-27 (a), the inverse voltage is about double the peak of the a-c input voltage as in the half-wave rectifier. Thus the inverse current is increased and the output voltage is reduced in this circuit.

To counteract this effect, metallic-oxide rectifiers are usually used in a bridge rectifier type of circuit because the small inverse voltages cause a minimum of reduction in the output voltage on the non-conducting half cycle of each pair of units.

In this same connection, copper-oxide rectifiers are frequently used without any input step-up transformer because of their low inverse peak voltage rating. The bridge circuit is particularly useful in such cases since it provides full-wave rectification without needing a transformer with a centre-tapped secondary. A bridge circuit using a metallic-oxide rectifier is shown in Fig. 3-28 (a). In Fig. 3-28 (b) and (c) the same circuit is redrawn to show how the rectifier may be built as a single unit.

The copper-oxide rectifier is capable of producing only a low-voltage output. For this reason it is used in d-c supplies to operate control circuits and relays, to charge batteries, to run small d-c motors and to supply bias to grid circuits.

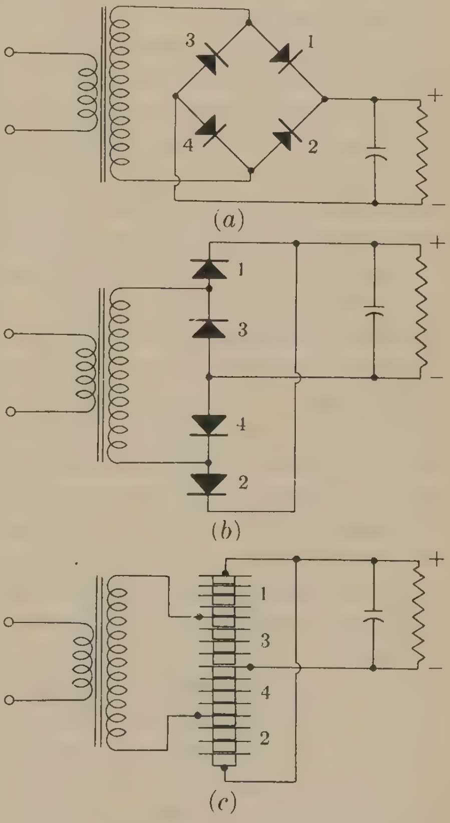


FIG.3-28 Bridge circuit employing metal rectifiers.

The current supplied may be quite large if the plates of the rectifier and fins for cooling are large. The selenium rectifier is slightly more efficient than a copper-oxide rectifier since its resistance in the conducting direction is slightly less and the unit necessary for the same current and voltage output is smaller. This latter fact is true because the selenium-coated plates can withstand a higher inverse voltage than corresponding copper plates and the operating temperature may be higher without damage. The selenium rectifier is sometimes used in a power supply to provide the anode voltages. The copper sulphide rectifier is characterized by small size and weight, by its ability to operate at a high temperature, good voltage regulation and definitely limited life. It should be used only for intermittent service.

Metal rectifiers are used because they give a reliable output, are not easily damaged except by corrosion of the metals used, require no maintenance and can withstand large overloads for short periods of time. In addition, provided they are not damaged by overloading, their life should be much longer than that of a valve rectifier.

3.8 Hot-cathode mercury-vapour rectifiers. The hard diode has a low efficiency as a rectifier because of the high anode resistance caused by the space charge around the cathode. This causes the voltage drop across the valve to increase as the current is increased with consequent loss of power and poor voltage regulation in the power supply. For low-power rectification, that is the development of d-c voltages up to about $400 V$ at several hundred milliamperes, hard diodes are completely satisfactory and are even advantageous because they can withstand more abuse. However, for power rectification, the *hot-cathode mercury-vapour rectifier* (Sec. 1.26) is used because it has much higher efficiency and better voltage regulation. The valve contains mercury vapour in equilibrium with liquid mercury. When the anode is at a positive potential of about $15 V$ with respect to the cathode, the electrons from the heated cathode are drawn to the anode with sufficient velocity to ionize the mercury vapour and a blue glow fills the tube. The positive ions are attracted to the cathode and neutralize the space charge surrounding it. Because of this fact the valve current may vary widely, even up to the peak emission current of the cathode and the voltage drop across the valve remains constant at about $15 V$. Because of this low voltage, the positive-ion velocities during normal operation are low and oxide-coated cathodes having large surface area are used in this type of valve. These give high emission for extremely low filament temperatures and the power used to heat the cathode is lower than in hard diodes. At the same time they are fairly fragile and care must be taken in the operation of the valve.

A comparison of the $e_b - i_b$ characteristics of a mercury-vapour diode and a high-vacuum diode is shown in Fig. 3-29. For very small voltages the two conduct slightly. However when the voltage across the mercury-vapour diode reaches a value of about 15 volts the mercury vapour ionizes and the valve conducts heavily. After ionization the mercury-vapour diode has almost perfect regulation since the voltage across it remains constant for any current up to the maximum emission current of the cathode. In comparison, the voltage across the high-vacuum diode continues to increase as increasingly larger currents are drawn.

The power loss in a mercury-vapour diode is much lower than in a high-vacuum diode. Consider the two characteristics shown in Fig. 3-29. The power loss in a valve is the product of the current flowing through it and the voltage

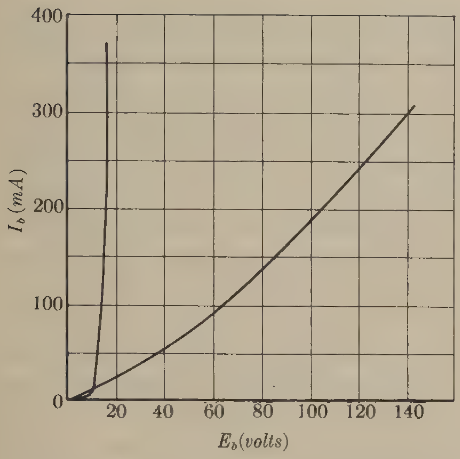


FIG.3-29 Comparison of $e_b - i_b$ characteristics of mercury-vapour and high-vacuum diodes.

across it. For the particular high-vacuum diode shown, a load of $300mA$ causes a voltage drop across the valve of $140V$. The power loss is therefore .3 x 140 watts or 42 watts. In the mercury-vapour diode the voltage is only $15V$ for the same current and the power loss is .3 x 15 watts or 4.5 watts. Thus the power loss is much less in the mercury-vapour diode.

The mercury-vapour diode possesses the advantages of higher thermal efficiency, better voltage regulation, lower filament power and higher output power than the high-vacuum diode. However, its use is limited by the fact that it is much more susceptible to damage and the operating conditions must be controlled within more exact limits. Some of the precautions which must be taken in the operation of a hot-cathode mercury-vapour rectifier are discussed in the following paragraphs.

If the anode voltage is applied before the cathode has had time to warm up properly, complete ionization does not occur and a higher voltage drop appears across the valve. If the voltage drop exceeds $22V$, the positive ions are attracted to the cathode so rapidly that the bombardment destroys the oxide-coated surface. This must be prevented by turning on the L.T. supply to the filaments at least half a minute before the anode voltage is applied. In some cases the operator must know and follow this procedure and in others an automatic-delay switch is used to prevent any possibility of damage to the valves. The delay in starting a set having mercury-vapour diodes constitutes a disadvantage in service equipment. If the set must be operated without delay it is never turned completely off but is left with filament circuits on at all times. This position is called the *stand-by* position and permits immediate operation at any time. However this advantage is obtained at the expense of the power used in heating the filaments when the set is not in operation.

The mercury-vapour diode must be protected against overloads since the current through the valve is limited only by circuit resistance and not by any compensating action of the valve. Overloads may be of two types, either a momentary short circuit on the load or a steady current which exceeds the maximum average anode current rating of the valve. If the peak current through the valve exceeds the possible emission current of the cathode for even an instant, as for example during a short circuit, the voltage across the valve rises rapidly and cathode disintegration results. If the average current is too large, the valve overheats and this greatly shortens its life.

As a protection against overloads, overload relays are usually inserted in the primary circuit of the power supply and in the H.T. return line. Proper protection is obtained when the following relays are employed:

(i) instantaneous overcurrent relays in the primary supply line which, in a three-phase system, are placed in two of the three-phase leads to insure operation when any one phase of the primary is overloaded

(ii) an instantaneous overcurrent relay in the grounded side of the output d-c circuit to operate at an instantaneous value of current somewhat below the rated maximum peak current

(iii) a time-delay overcurrent relay in the grounded side of the output circuit to operate on continued overload or to protect the valve from passing current in excess of its maximum average current rating for any appreciable period of time. The peak inverse voltage rating of the mercury-vapour diode is rather low. This is due to the presence of hot mercury vapour between cathode and anode. If on the negative half cycle, the voltage at the anode is too large, the mercury vapour ionizes and there is an arc-back from anode to cathode. A very large current flows during this time and the valve quickly overheats and is destroyed.

If, in a bank of mercury-vapour rectifier valves, one valve fails even momentarily in the inverse direction, it discharges the filter capacitors and the resulting excess current through the other valves may sputter active cathode material upon the anodes of these valves, causing some of them to arc back. If the offender can be eliminated immediately, probably no permanent damage will result in the other tubes. This however is not always a simple task without the aid of an arc-back indicator. It is recommended that an arc-back indicator be installed in series with each anode lead to identify the faulty tube in the event that arc-back difficulties are encountered.

For optimum operation of these valves the *ambient* temperature, that is the temperature of the air surrounding the valve, must be controlled. The valves should be spaced at least 6 inches apart to allow free circulation of air and it is sometimes necessary to apply a forced draft to keep the ambient temperature down. If the valves are closer than 6″, heat radiation from one may be absorbed by the other, raising the mercury temperature to a point beyond the safe limit even though the ambient temperature is within the specified range. The temperature must not become too high because the hotter the mercury vapour, the lower is the permissible peak inverse voltage and therefore the danger of arc-back in the valve is increased. If the ambient temperature becomes too low the mercury vapour in the valve does not ionize completely and cathode disintegration results. This usually occurs only when the set is first turned on. Frequently thermostatically controlled heating resistors are used in service equipment to keep the ambient temperature at about 75° when the supply is not in operation. This circuit is operated by the ships mains.

Precautions in installation are necessary because of the nature of the construction of the mercury-vapour diode. The valve must always be mounted in an upright position. This prevents the filaments from sagging as well as preventing the mercury from depositing on the active elements. These deposits may occur during periods of disuse and may cause arc-over if they are not eliminated. This process of evaporating the liquid mercury which may be deposited on the active elements of the valve, is called *conditioning* of a mercury-vapour diode.

Conditioning mercury valves is done after a period of disuse of a valve and when it is new. The rated filament voltage is applied for at least 15 minutes without any anode voltage. This distributes the mercury properly in the valve. The anode voltage obtained from a variable auto-transformer (variac) is a small portion of the operating voltage.

The rectifier is operated in this way for about 20 minutes during which time the anode voltage is gradually increased until the normal operating voltage is reached. If the equipment does not permit this procedure, the full anode voltage should be applied intermittently until the valve operates normally. If a new valve is not put into immediate use it should be stored in an upright position and the conditioning process repeated at least every three months. Some service equipment contains special conditioning circuits for its mercury-vapour valves so that spare valves may be kept ready for immediate use. When mercury-vapour valves are used a large amount of high-frequency current *transients* are generated. These are caused by the sudden ionization and de-ionization during the operation of the valves which results in current pulses which rise and fall very rapidly. An analysis of a current waveform of this type shows that it contains components at many frequencies. These components are radiated by power-supply leads and may be conducted into circuits where they cause unwanted noise or r-f *hash.* To help prevent this effect, r-f chokes acting as *hash suppressors* are inserted close to the valve in the anode leads of the power supply. Also the power supply is built in a separate chassis which is well shielded to prevent radiation of hash.

The hot-cathode mercury-vapour rectifier is widely used in transmitter power supplies and wherever high power and high efficiency are necessary. A mercury-vapour-rectifier power supply is shown in Fig. 3-30. The switch S_1 is the filament on-off switch which connects the filament voltage through the transformer T_1. Switch S_2, which may be closed 30 seconds after S_1 is closed, connects the high voltage to the valves through the H.T. transformer T_2. The circuit is a full-wave rectifier with choke-input filter L_3 and C_1. The resistances R_1, R_2 and R_3 form a bleeder network across the output terminals. The r-f hash produced in the circuit is eliminated by L_1 and L_2, the hash suppressors in the anode leads of the valves. Protection against overload on the valves is afforded by E_1, the overload relay. All the power supply current flows through E_1. The normal amount of current is not enough to open the contacts but if excessive current is drawn by the load, the magnetic field of E_1 opens the contacts which break the primary circuit

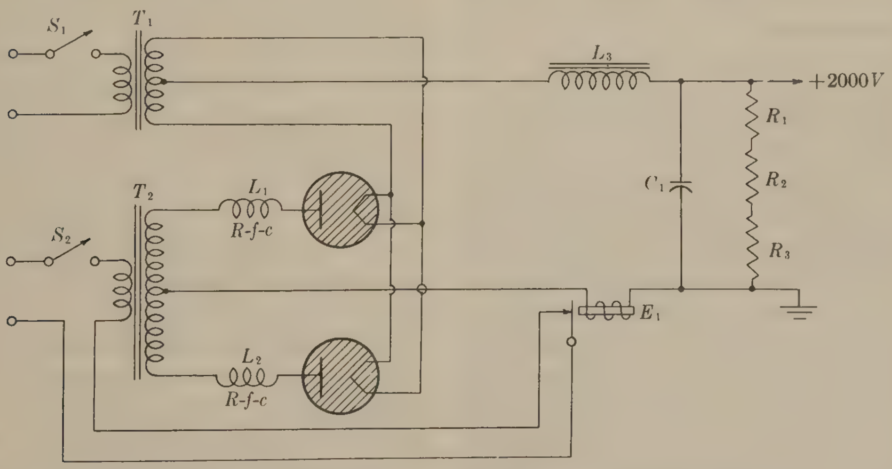


Fig.3-30 A mercury-vapour-rectifier power supply for use with a transmitter.

of T_2. Before operation can again be resumed, the contacts of E_1 must be reset. This is done only after the cause of the overload has been found and remedied.

3.9 Voltage multipliers. H.T. supply voltages are usually higher than the peak value of the a-c source. For this reason it is common practice to use a step-up transformer to increase the a-c voltage before rectifying it. However, if the required H.T. voltage is so high that it is not feasible to build a transformer to supply the full voltage or if it is desired to obtain an anode voltage suitable for operating small valves directly from the a-c source without using a transformer, then *voltage multipliers* are very useful.

The basic voltage multiplying circuit is the *voltage doubler* in which two capacitors are charged on alternate half cycles and arranged so that the voltages add together in the output. Such a circuit produces an output voltage which is close to twice the peak of the applied a-c voltage. This circuit may be connected in series with other rectifier circuits to produce voltages three, four or even more times the peak value of the input voltage. The operation of these circuits is discussed in detail in the following sections.

A. The Full-Wave Doubler

The circuit of a full-wave voltage doubler which is the conventional type of doubler circuit is shown in Fig. 3-31. When the end A of the transformer secondary is positive with respect to the end B a voltage is applied across V_1 and C_1 in series which makes the anode of V_1 positive with respect to its cathode. Thus electrons flow from the upper plate of C_1, through V_1 and accumulate on the lower plate of C_1 charging the capacitor to the peak voltage of the applied sine wave with polarity as shown. During a negative half cycle B is more positive than A and V_1 is cut off while V_2 conducts and C_2 is charged to the peak of the applied voltage, making the upper plate positive with respect to the lower plate (Fig. 3-31). The output is taken across C_1 and C_2 in series and is therefore equal to the sum of the voltages on C_1 and C_2. If the load current is small, this is very nearly twice the peak of the sine wave voltage applied to the valves. The resistance R is usually connected across the output in order to allow the capacitors to discharge after the supply is turned off and thus prevent injury caused by the unwitting touching of the high voltage by personnel. The resistance of R is in the range from 10 to 20 megohms.

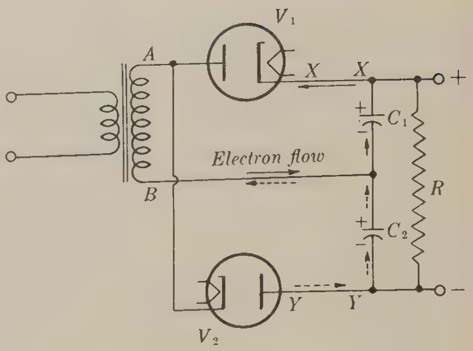


FIG.3-31 Full-wave voltage doubler circuit.

Fig. 3-32 shows the curves for the applied voltage, the d-c output voltage and the currents through V_1 and V_2. After C_1 and C_2 have been charged current flows through each valve only when the voltage across the valve is greater than the voltage across the capacitor which is in series with it. This occurs for only a short time during each half cycle and therefore the peak currents may be large. This circuit may only be used in applications where the average load current is not greater than a few milliamperes. If the current exceeds this, the capacitors discharge so much during each cycle that the output voltage drops considerably and in addition the peak current through the valves is very high. However a voltage doubler in which C_1 and C_2 are made large, about 10 microfarads, can supply a very large load current in pulses of a few microseconds duration without the voltage dropping noticeably. To prevent the valves from being damaged by the excessive peak currents which result from using large capacitors, resistors of a few thousand ohms resistance are placed in the circuit at the points labelled XX and YY (Fig. 3-31).

Since the ripple frequency in the output is twice the frequency of the input, this circuit is similar to the full-wave rectifier and is therefore called a *full-wave doubler*. Voltage-doubler valves, however, must have separate indirectly-heated cathodes for each diode section since the cathodes are at different potentials. This is different from the full-wave rectifier. For this reason twin diodes have been designed with separate cathodes to act as voltage-doubler valves.

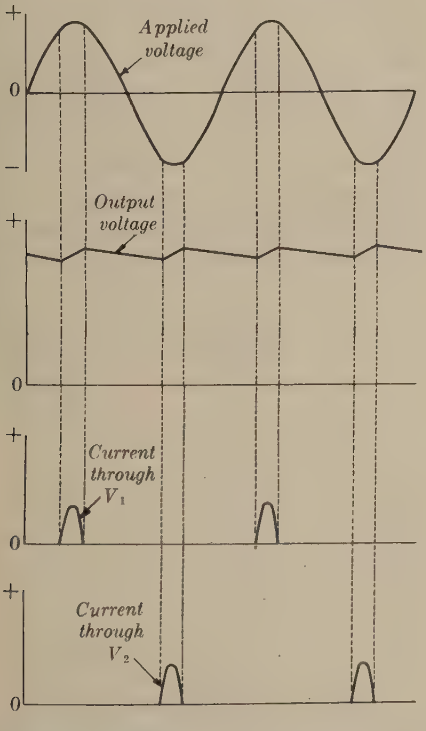


FIG.3-32 Waveforms of voltage and current in a full-wave doubler.

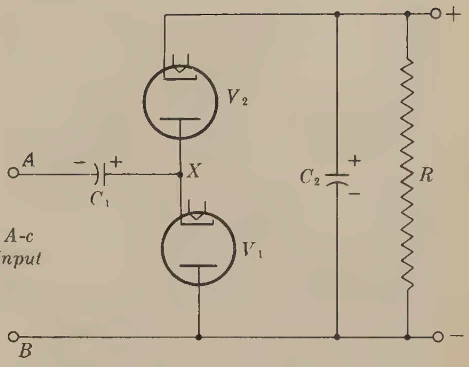


FIG.3-33 Half-wave doubler circuit.

B. The Cascade Doubler

Another type of voltage doubler is the *cascade doubler* shown in Fig. 3-33. During a negative half cycle, that is when A is negative with respect to B, the anode of V_1 is positive with respect to its cathode and V_1 conducts to charge C_1 with polarity shown. When A is at the negative peak the voltage across C_1 is equal to the peak of the a-c input less the small voltage across V_1 due to its anode resistance. Thus X is positive with respect to A by the peak voltage of the input sine wave, that is, it is at zero or ground potential. As the potential at A rises, point X becomes more positive. Since X always remains positive to A by the voltage across C_1, when A reaches its positive peak point X is positive with respect to ground by twice the peak input voltage. The graphs of the input voltage and the corresponding voltage at X are shown in Fig. 3-34 (a) and (b). During the positive half cycle of the input voltage V_1 is cut off and V_2 conducts, allowing C_1 to discharge and C_2 is charged up to the peak voltage between X and ground. Thus the output voltage taken across C_2 is very nearly twice the peak voltage of the input sine wave. On the next negative half cycle

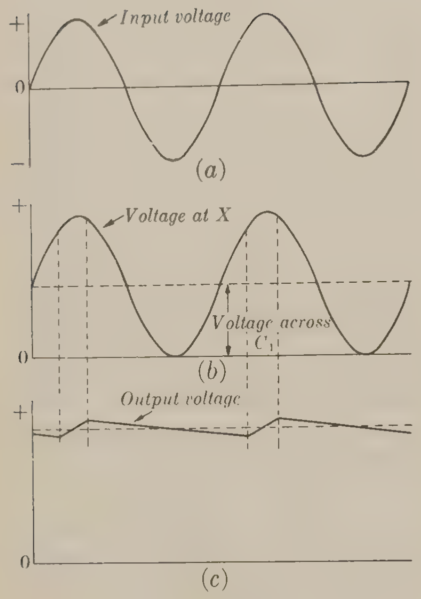


FIG. 3-34 Voltage waveforms in half-wave doubler.

the charge on C_1 is replaced through V_1 and C_2 keeps the current flowing through the load. As before the average load current may not exceed a few milliamperes so that only a small portion of the charge in the capacitors is lost when the charging currents are not flowing. This is necessary to keep the output voltage at its peak value and also prevent excessive ripple in the output. Fig. 3-34 (c) shows the waveform of the output voltage.

Another name for this circuit is the *half-wave doubler*. It is called this because V_2 acts as a half-wave rectifier of the voltage whose peak value is doubled by the action of C_1. The ripple frequency is the same as that of the source. For this reason this circuit is less desirable than the full-wave doubler since there are the same number of components and since the percentage ripple is greater. The cascade doubler however finds its application in multiplier circuits which give more than double the voltage output.

C. *Higher-Voltage Multipliers*

If a cascade doubler is combined with a half-wave rectifier an output voltage of three times the peak voltage of the input may be obtained. In Fig. 3-35 V_1 and V_2 together with C_1 and C_2 comprise the cascade doubler giving an output of $2E_{max}$ across R_1. V_3 is the half-wave rectifier producing an output of E_{max} across C_3 and R_2. These two outputs add up to $3E_{max}$ across R_1 and R_2.

In Fig. 3-36 a multiplier circuit is shown which uses two cascade doublers in series to produce an output voltage of almost 4 times the peak value of the input voltage.

3.10 Voltage regulators. It is shown in section 3.3E that the voltage output of a rectifier does not remain constant for different loads. Even when a choke-input filter is used with the best regulated valves, there is an IR drop in circuit components which increases with increased load current and thus changes the output voltage.

However, changes in output voltage across the load may take place even when the load remains constant. These changes are caused by variations in the input a-c voltage to the rectifier. Circuits which are designed to counteract the effect of the variations in load or

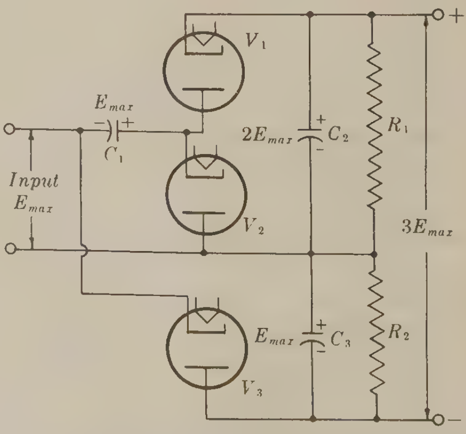


FIG. 3-35 Voltage tripler circuit.

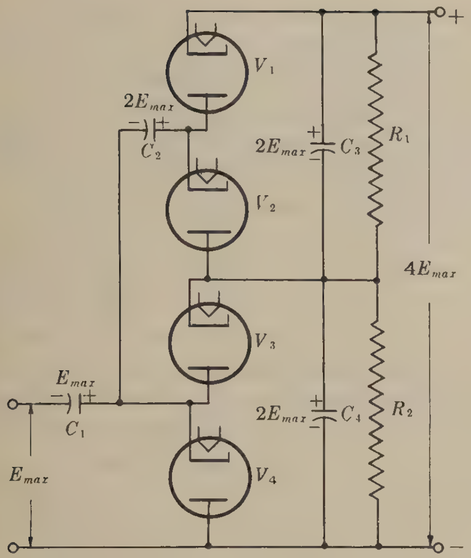


FIG.3-36 Voltage quadrupler circuit.

input a-c voltage on the output voltage of the rectifier are called *voltage regulators*. All voltage regulator circuits are based fundamentally on the concept of a voltage divider which consists of a variable resistor in series with the load (Fig. 3-37). If the output voltage drops, R is decreased to allow more current to flow and thus counteract the voltage drop across the load, or in other words, the voltage drop across R is decreased so that the voltage across the load remains unchanged. If the output voltage rises, then R is increased to cause a greater voltage to appear across it and thus keep the voltage across the load the same.

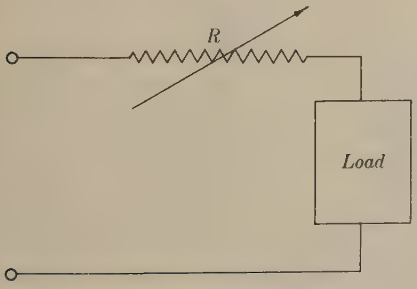


FIG.3-37 Circuit illustrating the principle of voltage regulation.

Instead of manually adjusting a variable resistance to suit the changing conditions, valve circuits and other devices have been designed to produce the effect of a variable resistance which automatically adjusts itself by the correct amount to counteract any changes in voltage across the load within very small limits. Some circuits only regulate the voltage when the changes are caused by changes in load and others only when the changes are caused by variation in the input a-c voltage but most circuits regulate the voltage in either case. The various types of circuits are described in the following sections.

A. *The Gas-Valve Regulator*

The *gas-valve regulator* is a diode valve filled with an inert gas and having a cold cathode. The voltage across the valve remains practically constant for a certain range of currents through the valve. This occurs because the internal resistance of the valve depends on the degree of ionization of the gas in it, and this changes considerably with very slight changes in anode voltage. Thus if the voltage rises slightly, the degree of ionization of the gas increases, the valve resistance decreases and the current increases. Over the operating range of the valve the product (IR) of the current and the internal impedance of the valve remain practically constant. For this reason it is called a *voltage regulator* or *VR valve.*

To act as a voltage regulator, the valve is connected in parallel with the load and a resistor R in series with the *VR* valve forms the other part of the required voltage divider as in Fig. 3-38. For controlling variations in output voltage caused by changes in the load, the action of the circuit is as follows. The total current in the circuit flows through R and then divides between the regulator valve and the load. If the load

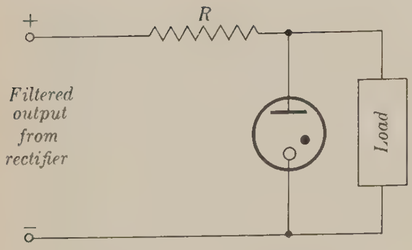


FIG.3-38 VR valve voltage regulator.

draws an increased current, i.e. the load resistance decreases, the voltage across the VR valve tends to decrease. The very slight decrease in voltage produces less ionization in the VR valve so that less current flows through it and the total current through R remains constant. Similarly if the load current decreases, the voltage across the VR valve tends to rise. This causes an increase in current through the valve which exactly counteracts the decrease in load current. The valve thus acts in such a way as to draw a constant current from the source through R and consequently the voltage across R and across the load remains constant. A constant output voltage is maintained over the range of current changes possible in the VR valve.

Changes in voltage from the rectifier filter are also absorbed by the VR valve and do not affect the output voltage across the load. If the rectifier output voltage tends to increase the voltage across the load and VR valve tends to increase, the VR valve then conducts more current which flows through R. This causes the voltage drop across R to increase and thus absorb the change. The voltage across the load is then kept constant.

The resistor R is necessary in the circuit to limit the excessive currents in the VR valve which may flow if the current to the load is cut off. The total current then flows through the valve and is limited by the size of R. The value of R is chosen so that the current through the valve has a safe value when there is no load current.

Only one value of voltage output can be obtained with any particular valve. If a different value of voltage is to be regulated, a different VR valve must be employed. Commonly used valves are the VR-75-30 and VR-150-30 which can regulate voltages of $75V$ and $150V$ respectively. If it is desired to regulate higher voltages two or more valves are connected in series as in Fig. 3-39. Care must be taken that the peak output voltage from the rectifier is large enough to cause the gas in the valve to ionize. This is about 30% greater than the constant voltage across the regulator when it is conducting. The variations of current that can be absorbed by one valve are relatively small, being about 5-30mA. When two valves are used in parallel, a current of twice the value can be regulated. However, this type of stabilization is used only in circuits which require a small current. A *voltage regulation* of about 1% can be obtained by the use of the VR valve.

Cold-cathode gas diodes may also be used to regulate the a-c input voltage to a transformer. Two neon valves are placed back to back across the a-c supply to the primary, and a resistance R is placed in series with the line, Fig 3-40 (a). Suppose it is desired to regulate a $110V$ supply voltage. Valves are chosen which ionize at, say, $105V$. Thus the

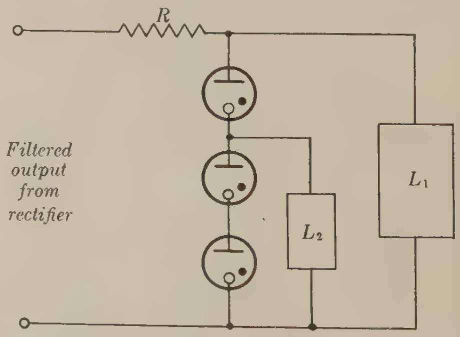
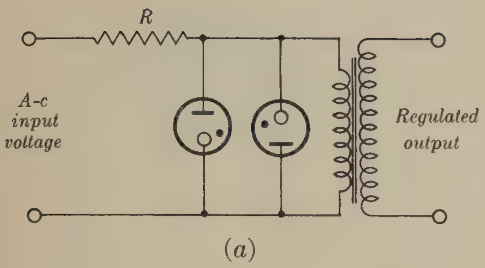


FIG.3-39 Use of several VR valves in series to stabilize a high voltage.



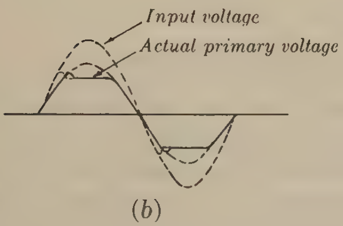


FIG.3-40 Use of cold-cathode gas valves to stabilize an a-c voltage.

input voltage may vary from 105 V upwards and the voltage applied to the primary will never rise above 105 V. This occurs because the neon valves ionize as soon as the voltage reaches 105 V and draw a larger current through *R*. Thus the increase in voltage is absorbed by *R* and the voltage across the primary remains at 105 V. This is illustrated by the graphs in Fig. 3-40 (b). Of course, this system is of no value if the input falls below 105 V, but it must be remembered that all voltage regulators are of value only within a limited range of variations.

B. The Hard-Valve Regulator

A triode valve capable of carrying the d-c current required by the load may act as a voltage regulator. This valve is inserted in series with the load as in Fig. 3-41. A triode valve has a variable anode resistance which is controlled by the grid voltage. This resistance is varied in such a way as to keep the voltage across the load from changing appreciably.

In Fig. 3-41 the grid potential is negative with respect to the cathode by the amount of the voltage *E* across the upper part of *R*. The position of the tap on *R* is adjusted until the correct voltage appears across the load. At this position the voltage output of the rectifier divides in the correct ratio across the valve and load in series. If the load resistance decreases, the voltage across the load tends to decrease. However, when a small decrease in voltage appears across the load and *R*, the voltage between the grid and cathode of the triode decreases. A large decrease in the resistance of the valve takes place when this bias decreases. Since the anode resistance decreases, the fraction of the rectifier output voltage appearing across the valve decreases and the voltage across the load is maintained at its initial value.

The vacuum-valve regulator also reduces the effect of changes in rectifier output voltage caused by variations in the input a-c voltage. If this voltage increases, a corresponding increase appears across *R* to apply more bias to the triode and thus increase its resistance. This increases the voltage across the valve which compensates for the increase in voltage across the load. Thus the load voltage remains the same.

This circuit is limited in its action. When the load current increases the grid voltage should be maintained at a constant lower value in order to hold the resistance of the valve at a decreased

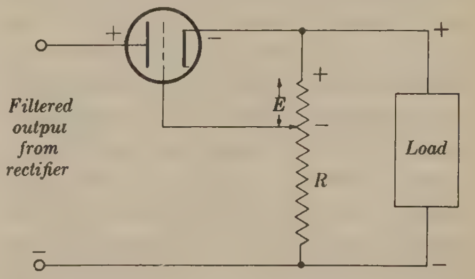


FIG.3-41 Simple hard-valve voltage regulator.

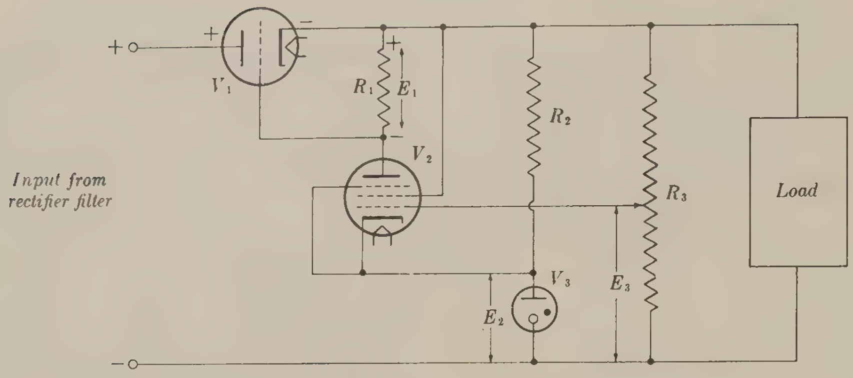


FIG.3-42 A more sensitive hard-valve voltage regulator.

value. However, after the load voltage is restored to its original value, the bias on the valve is increased slightly, the anode resistance and voltage across the valve increase causing the load voltage to increase slightly. Consequently the voltage across the load must continually decrease and increase by small amounts to obtain the desired action of the triode.

A more efficient method of controlling the resistance of the triode is shown in Fig. 3-42. In this circuit the change in load voltage necessary to effect a change in the resistance of the triode is very small because of the high amplification factor of the pentode valve. The action of the circuit is as follows. The grid voltage of the pentode is set by R_3 and controls the current in the pentode. The pentode current flowing through R_1 produces the bias on the triode. A setting of R_3 is made which gives the correct bias on the triode for the desired anode resistance, anode voltage and output voltage. When this is set, the small variations in output voltage, caused either by changes in load current or by changes in input voltage, are fed to the grid of the pentode. If the voltage across the load tends to decrease, the grid voltage on the pentode decreases causing less current through R_1. This decreases the bias E_1 on the triode and reduces the resistance of the triode. The voltage across the load is increased by this action to almost the same value it had previous to the change. As a result the change in load voltage is very small.

The VR valve is used to supply a constant positive voltage E_2 to the cathode of the pentode. This is necessary for stable operation of the pentode. R_2 is the series resistor which conducts the minimum current necessary for the VR valve to remain ionized. The position of the tap on R_3 makes E_3 a positive voltage less than E_2. Thus the pentode control grid is lower in potential than the cathode by the voltage $E_2 - E_3$.

A circuit of this kind is used for higher voltage and current outputs than are possible by the use of VR valves alone. The output voltage is for all practical purposes a steady d-c. Since the output voltage is constant with varying input voltage from the rectifier, this circuit may also take the place of part of the ripple filter of the power supply.

C. The Carbon-Pile Regulator

A *carbon-pile regulator* prevents changes in a-c input from being applied to the rectifier circuit.

A carbon pile is a series of carbon plates pressed in contact with one another. If the pressure on the plates is increased the resistance of the pile decreases; if the

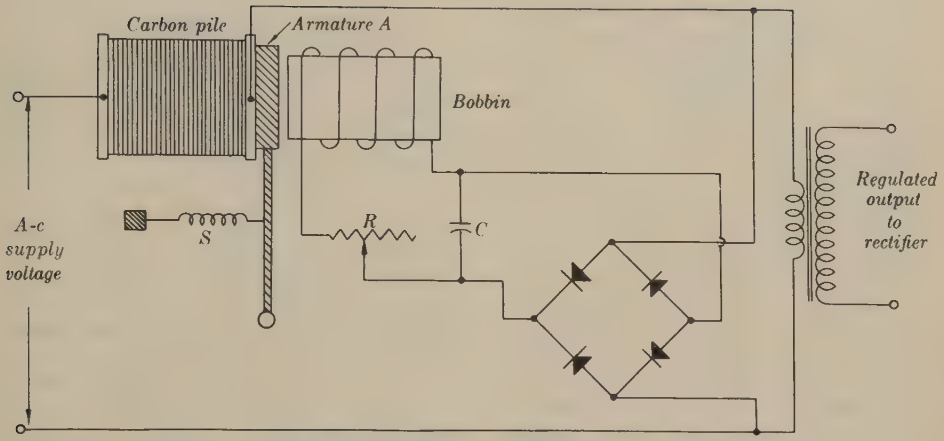


FIG. 3-43 Carbon-pile regulator.

pressure is decreased the resistance increases. By applying a variable pressure the resistance of the carbon pile is made to vary in such a way that the output voltage is kept constant with changes in input voltage.

The circuit diagram of a carbon-pile regulator is shown in Fig. 3-43. It is usually used in the primary circuit of the transformer of a rectifier power supply. The carbon pile is connected in series with the primary winding so that the total input current passes through it. A spring *S* applies pressure to the armature *A* which in turn compresses the plates. The magnetic field of the bobbin controls the pressure of the armature on the carbon pile. The current for the bobbin is supplied by a metallic-oxide bridge rectifier which is in parallel with the primary winding of the transformer. The variable resistor *R* in the rectifier output is an external means of varying the current in the bobbin, which changes the pressure of the armature on the carbon pile and thus sets the output voltage. The a-c output is kept nearly constant at the value which is initially set by *R*. The action is as follows. If the input voltage rises, a greater current is produced by the bridge rectifier. This increases the current through the bobbin and produces a stronger field which lessens the pressure on the carbon pile. The resistance of the pile is increased and more of the voltage drop appears across the carbon pile. Thus the output voltage rises only slightly. This small rise is necessary to maintain the lessened pressure on the pile. For decreases in the a-c supply voltage the action is exactly the reverse.

D. Amperite or Ballast-Valve Regulator

A *ballast valve* consists of an iron wire enclosed in an envelope filled with hydrogen. It depends for its regulating action on the fact that the resistance of the wire changes with temperature. With large currents the wire becomes hot and the resistance increases; with small currents the reverse is true. The ballast valve is specially constructed so that over a certain range of voltage the resistance changes in such a way that the current remains practically constant. Thus the valve may be used as a *constant-current device* or as a voltage regulator.

As a voltage regulator it may be used to counteract changes in the input a-c voltage. It may be placed in the primary of the transformer or in the output of the rectifier following the filter (Fig. 3-44). The action may be described as follows.

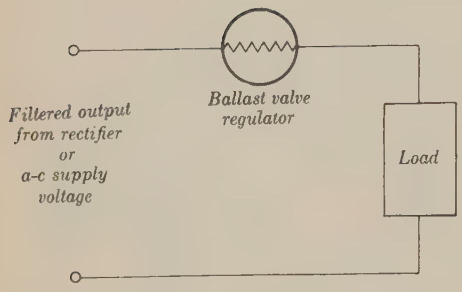


FIG.3-44 Amperite or ballast-valve regulator.

If the voltage increases, the current tends to increase and the wire heats up. The increased resistance in the ballast valve absorbs the increase in voltage and thus the voltage across the load remains the same.

Note that this device does not regulate the voltage if the load changes. If the load resistance decreases, more current is drawn from the rectifier and the output voltage falls. In addition, the greater current causes the resistance of the ballast valve to increase and the load voltage is decreased even further by the additional drop across the valve.

The ballast valve, however, is used most commonly as a constant-current device. It is usually placed in series with several elements through which it is desired to maintain a constant current, such as a circuit in which the filaments of several valves are connected in series. Fig. 3-45 shows the characteristic of a typical ballast valve and illustrates its constant-current property.

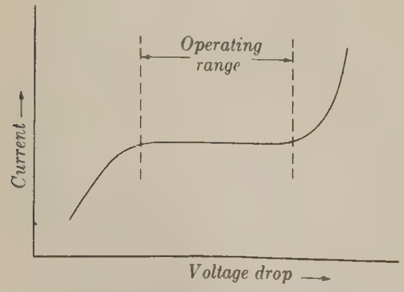


FIG.3-45 Characteristic curve of a ballast valve showing its constant-current property.

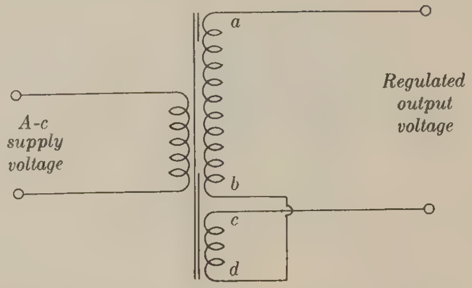


FIG.3-46 Saturated-core transformer regulator.

E. Saturated-Core-Transformer Voltage Regulator

Changes in the primary voltage of a transformer cause smaller changes in secondary voltage when the core is saturated than when the core is not saturated. This levelling off of output voltage after saturation is used in stabilizing the output voltage of a transformer for changes in input voltage.

In the transformer (Fig. 3-46) the upper part *ab* of the secondary is wound on a thin core which saturates at the output voltage desired. This part of the secondary has many turns and the graph of its output E_{ab}, with changing input voltage is shown in Fig. 3-47. The lower part *cd* of the secondary has few turns and is wound on a normal core that does not saturate in the voltage range desired. Thus, its output E_{cd} is proportional to the input voltage as shown. The secondaries *ab* and *cd* are connected so that their outputs oppose each other and therefore the output voltage is the difference between E_{ab} and E_{cd}. The number of turns in the sections *ab* and *cd* are chosen so that the output voltage does not change appreciably beyond saturation over the range of input voltages desired.

F. A Current Regulator

A *current regulator* is a circuit which causes a constant current to flow in a load regardless of changes in input voltage or load resistance. This circuit,

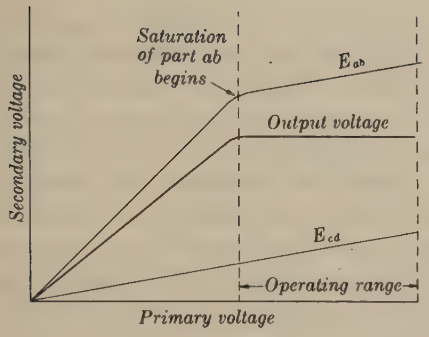


FIG.3-47 Graphs illustrating operation of the saturated-core transformer.

shown in Fig. 3-48, makes use of that property of a pentode which allows only small changes in anode current with large changes in anode voltage over the operating range *A* to *B* (Fig. 3-49). Since the load is placed in series with the pentode, the pentode current flows through it. To maintain the screen voltage constant for stable operation, a *VR* valve is connected between the screen and ground. The resistor R_1 is small enough that the regulator valve conducts and large enough to limit the screen current to a safe value when the anode current is small. The initial operating current of the circuit is set by R_2 which varies the bias applied to the grid. This is set for the desired current through the load.

The action of this current regulator is as follows. If the load tends to draw more current, the current through R_2 and R_3 causes a greater bias to be applied to the pentode. This increased bias decreases the anode current. The net increase in current through the pentode and load is very small. Thus, the anode current is nearly constant with changes in anode voltage over the operating range (Fig. 3-49).

This type of current regulator is capable of limiting the current change to the order of 1*mA* as the voltage across the circuit varies from 300 to 1000*V* or as the load resistance changes from 0 to 50,000 ohms.

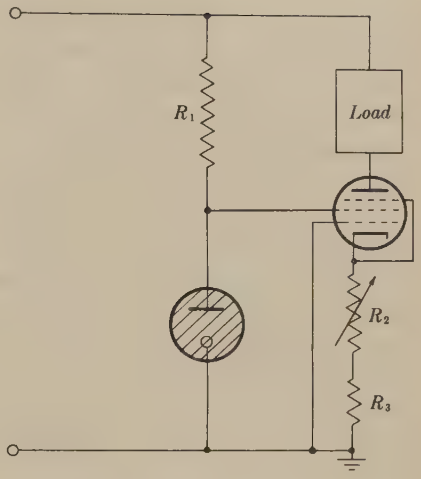


FIG.3-48 A current regulator circuit.

3.11 Vibrator power supplies. The *vibrator* is a device used to obtain large H.T. voltages for transmitters or receivers from a low-voltage d-c source. It is most commonly used in portable and mobile sets where 6*V* or 12*V* storage batteries are the only available source of power.

A vibrator is essentially nothing more than a simple *interrupter*, similar in many respects to a buzzer, which changes the d-c from a storage battery to an a-c. A battery connected directly to a transformer produces no secondary voltage because there is no changing field in the primary. By means of a vibrator the connections of the battery to the transformer are reversed at a rapid rate. The resulting alternating square-wave d-c pulses in the primary produce a changing

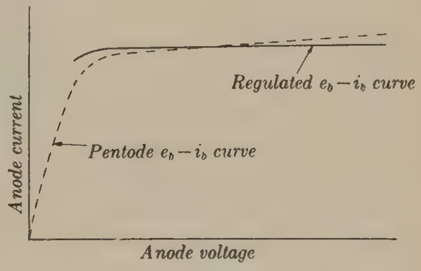


FIG.3-49 Characteristic curve of current regulator.

magnetic field which induces a high-voltage a-c in the secondary. This high-voltage a-c is then rectified and filtered to produce a smooth H.T. voltage suitable for receivers or transmitters.

There are two main types of vibrators in use. These differ only in the method of rectifying the high-voltage a-c. One type employs a valve rectifier in the secondary circuit and is called a *non-synchronous vibrator*. The other type rectifies its own output by using a second pair of vibrator contacts, thus eliminating the need for a valve and it is called a *synchronous vibrator*.

A. Non-Synchronous Vibrator Supply

The circuit of Fig. 3-50 shows a non-synchronous vibrator supply. When S_1 is open the vibrator reed rests in the position shown, mid-way between the contacts. If S_1 is closed current flows through the bobbin coil and the lower part of transformer T_1. The magnetic field of the bobbin pulls the reed down until it makes contact at *A*. This applies the full battery voltage to the bottom half of T_1 and shorts the bobbin out of the circuit. A large current flows in the bottom half of T_1 in the direction shown but since the reed is no longer pulled by the bobbin, it springs back quickly to break the contact. Its momentum carries it up to make contact at *B*. The battery is then connected to the top half of T_1 so that a large current flows in the direction shown. As soon as the reed leaves contact *A*, however, the bobbin is energized again and attracts the reed back, breaking the current in the upper half of the transformer. This action repeats itself many times a second to produce the a-c input for the transformer.

The voltage waveform in the primary of the transformer when C_2 is not in the circuit is shown in Fig. 3-51 (a). At *A* on the diagram the power contacts are closing on one direction of swing of the reed, connecting the primary of the transformer to the terminals of the battery. The contacts remain closed a length of time t_1 until point *B* where the reed has started its return swing and has opened this pair of power contacts. The reed now requires a length of time t_2 to continue this return swing to the point where the opposite pair of contacts close at *C* on the diagram. The current now flows through the primary of the transformer in the opposite direction. This second pair of contacts remains closed for the time t_3, after which the reed has reversed its direction again and has continued its return swing far enough to open these contacts at *D*. The reed then

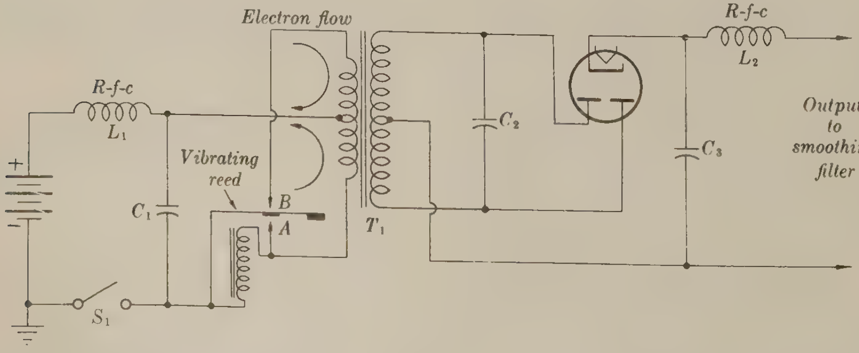


FIG. 3-50 Non-synchronous vibrator circuit.

requires a length of time t_4 to travel between the second set of contacts at D and the original set at A where the cycle ends and a new one begins.

Current can only flow from the battery while the power contacts are touching, i.e. during the time periods t_1 and t_3. A *timing* or *buffer* capacitor C_2 is required to protect the circuit during the time that the reed is moving from one set of contacts to the other, i.e. t_2 and t_4. If no capacitor is used an extremely high voltage is induced in the transformer because of the sudden stoppage of the current and the rapid collapse of the magnetic field. This causes severe arcing and failure of the vibrator. Further, when the contacts close at C the full battery voltage is applied across the contacts causing a spark to jump the gap just before the contacts close, which is also detrimental to good contact life. By connecting a capacitor across either of the transformer windings and selecting the right capacitance these conditions may be avoided. This capacitor acts as a *tank* in which energy is stored during the *contact* intervals, t_1 and t_3, and discharges through the transformer winding during the *off contact* intervals, t_2 and t_4, to supply energy to the transformer. Thus the current is maintained and the waveform takes the form shown in Fig. 3-51 (b) when there is no load on the secondary, and that of Fig. 3-51 (c) for a load. Fig. 3-51 (d), (e), (f), (g) shows the effects of faulty operation on the waveform. Fig. 3-51 (d) shows the effect of too small a timing capacitance when the vibrator is loaded and Fig. 3-51 (e) shows the effect of the same fault when the vibrator is not loaded. Fig. 3-51 (f) shows the effect if the vibrator is worn out or badly adjusted and is making poor contact on one side while Fig. 3-51 (g) shows the effect of bouncing contacts caused by worn-out contacts or a sluggish reed.

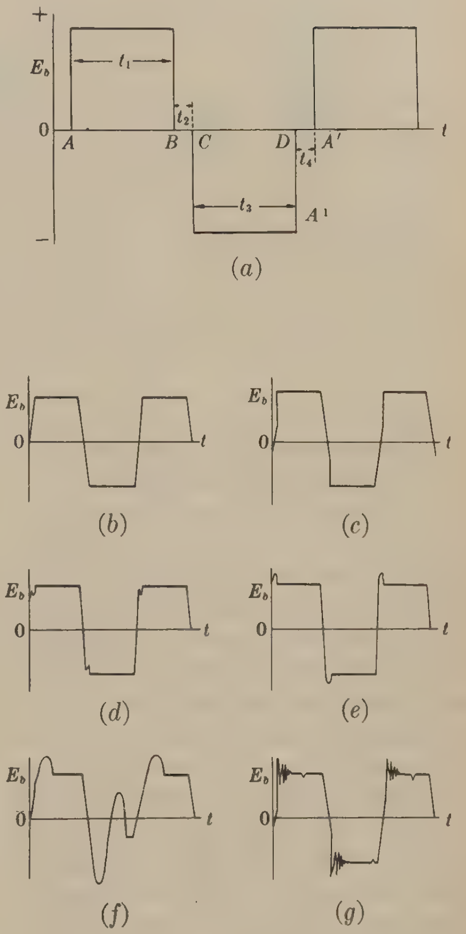


FIG. 3-51 Voltage waveforms at primary of a non-synchronous vibrator.

Sparking at the contacts with resultant r-f interference, commonly referred to as *hash*, is suppressed by the filter choke r-f-c L_1 and capacitor C_1 placed in the input leads. The entire unit is also placed in a metal can to shield nearby circuits from any interference caused by the vibrator.

The d-c output voltage is produced by a full-wave rectifier using a double diode valve. The r-f-c L_2 and capacitor C_3 comprise an L-section filter for r-f hash. The output from this circuit is passed on to a ripple filter which smooths the output to a steady d-c.

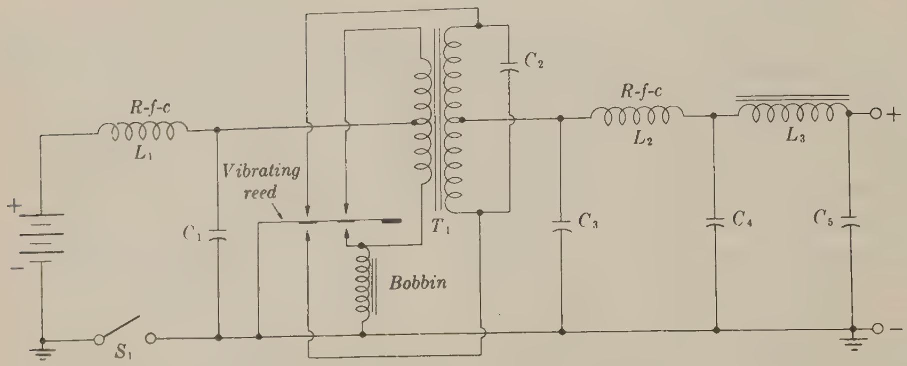


FIG.3-52 A synchronous vibrator circuit.

B. Synchronous Vibrator Supply (Self-Rectifying Type)

A *synchronous vibrator* supply circuit is shown in Fig. 3-52. The primary circuit is the same as in the non-synchronous type. An additional set of contacts on the vibrator reed is attached to the secondary circuit. This second set of contacts grounds one end of the secondary winding on one half cycle and the other end on the next half cycle. The output voltage to the filter is taken between the centre tap of the transformer and ground. Since the alternate grounding of the ends of the secondary is in step with the reversal of primary current, the polarity of the secondary centre tap is always the same relative to ground. Thus no rectifier valve is necessary to produce a d-c output.

The waveform of the primary voltage of a properly-adjusted synchronous vibrator operating under load is shown in Fig. 3-53. The small peaks shown at the beginning and end of time intervals t_1 and t_3 are normal and do not produce *hash* if they appear as shown. They are caused by the increased voltage drop in the primary circuit when the secondary load is connected, since the vibrator is adjusted so that the interrupter contacts close before and open after the rectifier contacts. This is accomplished by spacing the rectifier contacts slightly wider than the interrupter contacts and thus the load is broken at high voltage and low current rather than at low voltage and high current. The reason for this is that it results in less severe arcing and therefore in longer contact life.

3.12 R-f oscillator power supply for c.r.t. Although the anode supply for a cathode-ray tube requires a high voltage it requires very little power. The conventional transformer circuit used to obtain the high voltage is capable of producing much more power than is needed. It also has the disadvantages of being bulky, heavy and costly. Consequently another type of supply called an *r-f oscillator power supply* is used, especially with electrostatic deflection cathode-ray tubes, which produces a high voltage with the required amount of power. Since it consists only of two valves and an r-f transformer, its light weight and

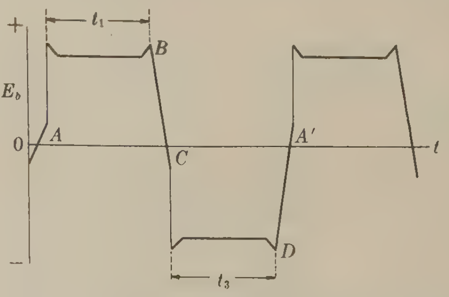


FIG.3-53 Voltage waveform at primary of a synchronous vibrator under load conditions.

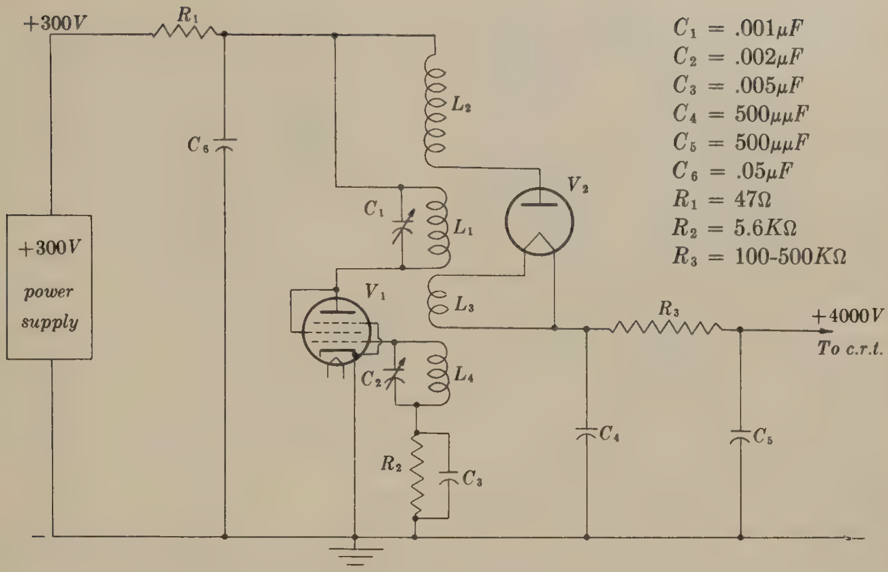


FIG.3-54 R-f oscillator power supply circuit.

compact size are further advantages and in addition its output is independent of the deflection system of the c.r.t.

A circuit diagram of this power supply is shown in Fig. 3-54. The a-c for the rectifier is produced by an r-f oscillator using a power output type of valve, V_1, in the circuit. The 300V d-c supply used for the other cathode-ray tube circuits supplies power for the oscillator which is a parallel-feed tuned-anode oscillator operating at 50-500 Kc/s and developing 10-15 watts output. The tuned-anode circuit L_1C_1 is tuned to the natural resonant frequency of the high-voltage winding L_2 of a step-up transformer. This provides a minimum amount of loading on the oscillator circuit. After the r-f has been stepped up to several thousand volts by L_2, it is applied to the diode V_2, and rectified. The output of the rectifier is filtered by C_4, R_3 and C_5. Since the rectifier diode requires only about $\frac{1}{4}$ watt to heat the filament, the power for this is taken from the r-f oscillator by an additional one or two turns, L_3 on the secondary of the oscillator coil. This eliminates the need for a conventional transformer having high-voltage insulation. The position of L_3 is critical and must not be changed since there is no way of measuring the voltage and the filament can easily be overloaded and burnt out. A comparison of the brightness of this filament with that of other ordinary valves is one way of estimating whether the position of L_3 is correct or not.

This power supply has good voltage regulation and there is no need for an additional voltage regulator. The regulation curve is shown in Fig. 3-55. For

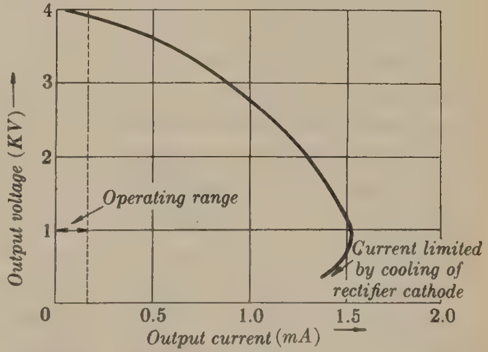


FIG.3-55 Regulation curve for r-f oscillator power supply.

variations in current of 0-200 microamps which occur as the electron beam is modulated, there is only a 5% fluctuation in voltage. If more current than this is drawn the voltage falls rapidly because loading reduces the Q of the tuned circuit so that the a-c voltage to the rectifier is reduced. Another factor in reducing the output when the supply is heavily loaded is the decrease in rectifier filament supply since it also gets its power from the oscillator. With a smaller heater current, less current flows through the rectifier.

The drop in voltage when the circuit is loaded has the advantage that the power supply is not as dangerous should a person inadvertently touch it during servicing of the equipment. The chances of a lethal shock are also reduced by making the filter resistor R_3 of the order of $100\,K\Omega$ to $500\,K\Omega$. However, the usual precautions taken when dealing with high voltages must still be observed; particularly, the connections to the rectifier valve must not be touched since a severe r-f burn may result.

3.13 Applied mathematics.

A. Full-wave Rectifier with Inductor Input Filter

1. D-C OUTPUT VOLTAGE.

In the inductor-input filter circuit shown in Fig. 3-56 (a) $E_{max} \sin \omega t$ is the a-c applied voltage to each valve. In this circuit e_i is the input voltage to the filter and load, i_i is the current to the filter and load which branches into i_c and i through the capacitor and load respectively. The effective resistance of a bleeder and load in parallel is represented by the resistor R and the inductor resistance by R_L.

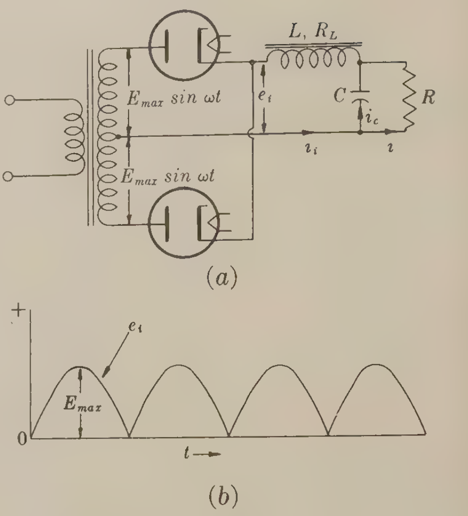


FIG.3-56 Circuit and waveforms of a full-wave rectifier with inductor-input filter.

In this circuit the inductor and capacitor action are such that under operating conditions the valves conduct during the complete half cycle when their anodes are made positive with respect to their cathodes by the voltage E_{max} sin ωt. The valve then acts as a short circuit which applies the input voltage directly to the filter and load circuit. This is true if the voltage drop across the valves is assumed to be zero. Since it is usually not a large value, this assumption is reasonably correct. Thus the voltage e_i is of the form shown in Fig. 3-45 (b) and is given by

$$e_i = E_{max}\left(\frac{2}{\pi} - \frac{4}{3\pi}\cos 2\omega t - \frac{4}{15\pi}\cos 4\omega t \ldots\right)$$

From this equation it may be seen that the d-c component is represented by the first term $E_{max}\frac{2}{\pi}$ and the a-c components by the remainder of the terms.

Thus

$$E_{d\text{-}c} \text{ across } R = \frac{2E_{max}}{\pi} \times \frac{R}{R+R_L}$$

and

$$I_{d\text{-}c} \text{ through } R = \frac{2E_{max}}{\pi} \times \frac{1}{R+R_L}.$$

2. A-C R.M.S. RIPPLE VOLTAGE IN THE OUTPUT.

The a-c ripple voltages occur at even multiples of the supply input frequency and are treated separately. The 2nd harmonic a-c voltage across the input is

$$e_2 = \frac{4E_{max}}{3\pi}\cos 2\omega t$$

The effective value of e_2 is

$$E_2 = \frac{1}{\sqrt{2}}\times\frac{4E_{max}}{3\pi}$$

∴ the effective value of i_2 is

$$I_2 = \frac{4E_{max}}{3\pi\sqrt{2}Z_i}$$

where Z_i is the input impedance of the filter and load computed at a frequency $2\omega t$.

The impedance Z_i is made up of $2\omega L$ in series with the combination of $\frac{1}{2\omega C}$ and R in parallel. However, in a filter $2\omega L \gg \frac{1}{2\omega C}$ and R in parallel with $\frac{1}{2\omega C}$ makes this a-c impedance still smaller. Consequently a good approximation of Z_i is given by $2\omega L$.

Thus $$I_2 = \frac{4E_{max}}{3\pi\sqrt{2}\times 2\omega L}$$

Since this is the second harmonic current passing through the filter, the voltage output across the load at this frequency is the same as the voltage across the capacitor.

$$\begin{aligned} E_2 &= I_2\times Z_c \\ &= \frac{4E_{max}}{3\pi\sqrt{2}}\times\frac{1}{2\omega L}\times\frac{1}{2\omega C} \\ &= \frac{4}{3\pi}\times\frac{E_{max}}{\sqrt{2}}\times\frac{1}{4\omega^2 LC} \end{aligned}$$

The above assumptions are even more exact for the fourth harmonic, i.e. $4\omega L \gg \frac{1}{4\omega C} \ll R$. E_4 is the r.m.s. value of the 4th harmonic of the input voltage appearing across the output and is calculated in a manner similar to that for the 2nd harmonic.

$$\begin{aligned} E_4 &= I_4\times Z_c \\ &= \frac{4}{15\pi}\times\frac{E_{max}}{\sqrt{2}}\times\frac{1}{4\omega L}\times\frac{1}{4\omega C} \\ &= \frac{4}{15\pi}\times\frac{E_{max}}{\sqrt{2}}\times\frac{1}{16\omega^2 LC} \\ &= \frac{1}{20}\times\frac{4}{3\pi}\times\frac{E_{max}}{\sqrt{2}}\times\frac{1}{4\omega^2 LC} \end{aligned}$$

This a-c component is only $\frac{1}{20}$ of the 2nd harmonic and can therefore be neglected.

3. PERCENTAGE RIPPLE IN THE OUTPUT.

The percentage ripple at the 2nd harmonic of the input frequency equals

$$\begin{aligned} &\frac{\text{r.m.s. a-c voltage output}}{\text{d-c voltage output}}\times 100 \\ &= \frac{\frac{4}{3\pi}\times\frac{E_{max}}{\sqrt{2}}\times\frac{1}{4\omega^2 LC}}{\frac{2E_{max}}{\pi}\times\frac{R}{R+R_L}}\times 100 \\ &= \frac{R+R_L}{6\sqrt{2}\omega^2 LCR}\times 100 \end{aligned}$$

For a supply voltage of 60 c/s this formula may be simplified considerably. $\frac{R+R_L}{R} = 1$ since R_L is usually small compared to R.

$$\begin{aligned} \text{Percentage ripple} &= \frac{100}{6\sqrt{2}\omega^2 LC} \\ &= \frac{100}{6\times 1.41\times(2\times 3.146\times 60)^2\times C\times 10^{-6}\times L} \\ &= \frac{100}{1.2LC} \end{aligned}$$

Thus percentage ripple $\doteqdot \frac{100}{LC}$

where

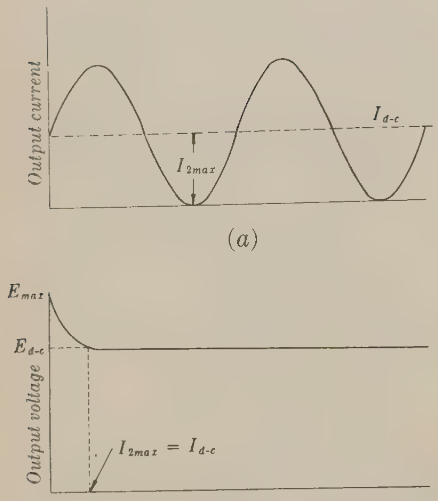
L = inductance of filter in henries,

C = filter capacitance in microfarads.

Since the 4th harmonic percentage ripple would be $\frac{1}{20}$ of the 2nd harmonic percentage ripple it may be neglected.

4 CALCULATION OF CRITICAL LOAD RESISTANCE R AND INDUCTANCE L.

The above calculations have assumed that current is flowing in the circuit at all times and that there is no transient condition. This, however, does not hold true for small values of d-c output current. The total current i_i is made up of the sum of two other currents: the d-c output $I_{d\text{-}c}$ and the second harmonic a-c I_2 neglecting the higher-harmonic a-c currents which are negligible. Consequently the current is a varying d-c as shown in Fig. 3-57 (a).



(a)

(b)

FIG. 3-57 Output voltage characteristic for a choke-input supply.

In the condition shown $I_{d\text{-}c}$ is equal to the peak value of I_2 and the current reaches zero at only one instant every cycle. If $I_{d\text{-}c}$ is greater than this value, the current does flow at all times and the previous calculations hold true. However, if $I_{d\text{-}c}$ is less than this value then the current stops at some time in the cycle and a transient condition occurs in which the capacitor discharge supplies the current for a short period. In this case, the capacitor gains enough charge in a shorter period than $\frac{1}{2}$ cycle to raise its voltage to the point where the valve is cut off. Consequently when $I_{d\text{-}c}$ is smaller than the peak of the second harmonic current, the voltage output rises. The output voltage characteristic for a choke-input supply is shown in Fig. 3-57 (b). In order to prevent operation on the rising part of the characteristic $I_{d\text{-}c}$ must be equal to or greater than $I_{2\,max}$. Since $I_{d\text{-}c}$ depends on the load and $I_{2\,max}$ depends on the value of the inductance, the critical value of load resistance and filter inductance can be calculated by equating the expressions for $I_{d\text{-}c}$ and $I_{2\,max}$.

$$I_{d\text{-}c} = \frac{2E_{max}}{\pi}\times\frac{1}{R+R_L}$$

$$I_{2\,max} = \frac{4E_{max}}{3\pi\times 2\omega L}$$

$$\therefore\ \frac{2E_{max}}{\pi}\times\frac{1}{R+R_L}\geq\frac{4E_{max}}{3\pi\times 2\omega L}$$

$$\therefore\ R+R_L \leqq 3\omega L$$

If $f = 60$ c/s, then

$$R+R_L \leqq 3\times 2\times 3.142\times 60\times L$$
$$\leqq 1131L$$

In most cases R_L is small compared to R and 1131 can be replaced by 1000 to arrive at an approximate minimum value of load resistance R_{crit} to draw the critical current with a given inductance L.

Thus, $R_{crit} \leqq 1000L$

This resistor is used as a load resistor across the output of the power supply to draw a minimum current for constant output voltage conditions. For a given load resistance, the critical inductance L_{crit} may be calculated from the same formula:

$$L_{crit} = \frac{R}{1000}$$

where

L_{crit} = critical inductance, in henries
R = effective resistance of bleeder and load in ohms.

5. D-C OUTPUT AND RIPPLE VOLTAGE USING A DOUBLE L-SECTION FILTER.

In the circuit of Fig. 3-58 is shown a filter consisting of two inverted L-sections. The ripple voltage for the first L-section has been calculated as

$$E_2 \quad \frac{4}{3\pi}\times\frac{E_{max}}{\sqrt{2}}\times\frac{1}{4\omega^2 L_1 C_1}$$

The voltage divides in the ratio of the impedances in the next section. Assuming Z_{C2} small in comparison with Z_{L2}, voltage across C_2 is $\frac{Z_{C2}}{Z_{L2}}\times$ voltage applied.

$$= \frac{1}{2\omega C_2}\times\frac{1}{2\omega L_2}\times\text{applied voltage}$$

$\therefore$ E_2 across C_2

$$= \frac{4\,E_{max}}{3\pi\sqrt{2}}\times\frac{1}{4\omega^2 L_1 C_1}\times\frac{1}{4\omega^2 L_2 C_2}$$

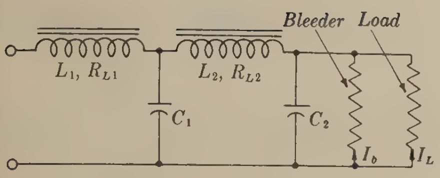


FIG.3-58 Double L-section filter.

Thus, the percentage ripple

$$= \frac{E_2}{E_{d\text{-}c}} \times 100$$

$$= \frac{\dfrac{4E_{max}}{3\pi\sqrt{2}} \times \dfrac{1}{4\omega^2 L_1 C_1} \times \dfrac{1}{4\omega^2 L_2 C_2} \times 100}{\dfrac{2E_{max}}{\pi} \times \dfrac{R}{R+R_{L1}+R_{L2}}}$$

$$= \frac{R+R_{L1}+R_{L2}}{24\sqrt{2}\,\omega^4 L_1 L_2 C_1 C_2 R} \times 100$$

The d-c voltage output may be calculated as follows. $E_{d\text{-}c}$, which is the input to the filter, is $\frac{2E_{max}}{\pi}$ where E_{max} is the peak value of the input a-c voltage. This voltage is dropped by the total resistance $R_{L1}+R_{L2}$ of the inductors. Since the d-c current in milliamperes is the sum of I_b, the bleeder current, and I_{load}, the load current,

$\therefore$ the voltage drop across the inductors is

$$\frac{(I_b+I_{load})}{1000} \times (R_{L1}+R_{L2}).$$

The voltage drop across the valve E_V must also be taken into account.

$\therefore$ $E_{d\text{-}c}$ is given by

$$\frac{2E_{max}}{\pi} - \frac{(I_b+I_{load})(R_{L1}+R_{L2})}{1000} - E_V$$

If the r.m.s. value of the input voltage E is known, then since $E_{max} = \sqrt{2}E$, $\therefore$ $E_{d\text{-}c}$ equals

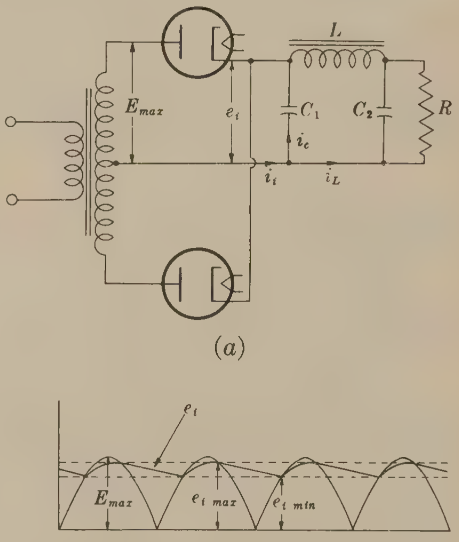
$$\frac{2\sqrt{2}E}{\pi} - \frac{(I_b+I_{load})(R_{L1}+R_{L2})}{1000} - E_V$$

$$= .9E - \frac{(I_b+I_{load})(R_{L1}+R_{L2})}{1000} - E_V.$$

B. D-C Output Voltage of a Full-Wave Rectifier with Capacitor-Input Filter

The behaviour of a capacitor-input filter cannot be derived from the waveforms of the inductor-input filter since the current is not flowing at all times through the valves. Thus it cannot be considered that the valve forms a short circuit which effectively connects the input voltage to the filter. The circuit and waveforms of a full-wave rectifier with capacitor-input filter are shown in Fig. 3-59 (a) and (b) respectively. In this circuit experimental results are usually used in determining the output. In the valve manual a graph of output voltages at various load currents is shown for nearly every valve used. Fig. 3-60 illustrates such a set of curves.

However, a certain mathematical analysis can be performed which predicts the performance of the above circuit fairly well. If the first capacitor is large and the valve can withstand the high peak current necessary, the voltage across the capacitor $e_{i\,max}$ approaches



(a)

(b)

FIG.3-59 Circuit and waveforms of a full-wave rectifier with capacitor-input filter.

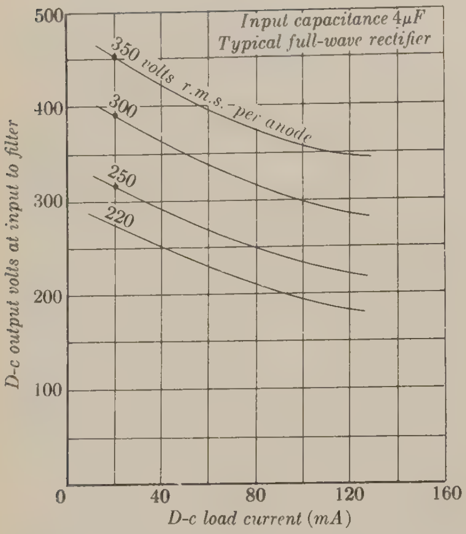


FIG.3-60 Output voltages at various load currents.

E_{max}. The amount by which $e_{i\,max}$ is less than E_{max} is designated by $\triangle e$.

$$\text{Hence } e_{i\,max} = E_{max} - \triangle e.$$

The value of the voltage across the capacitor after it has discharged through the L-section filter and load R is represented by $e_{i\,min}$. In this filter section since L and C_2 reduce the ripple current to a small value compared to the d-c current, the current may be assumed to be a steady d-c. Consequently the discharge of C_1 through L, C_2 and R is linear. The decrease in voltage is then given by

$$e_{i\,max} - e_{i\,min} = \frac{I_{d\text{-}c}t_1}{C_1}$$

where t_1 is the time interval over which a valve is not conducting in each half cycle of the input supply voltage.

Thus the d-c voltage across R is the average of $e_{i\,max}$ and $e_{i\,min}$ less the d-c voltage drop across L.

$$\therefore\ E_{d\text{-}c} = \frac{e_{i\,max}+e_{i\,min}}{2} - I_{d\text{-}c}R_L.$$

Since, from the above equation,

$$e_{i\,min} = e_{i\,max} - \frac{I_{d\text{-}c}t_1}{C_1}$$

or

$$\frac{e_{i\,min}}{2} = \frac{e_{i\,max}}{2} - \frac{I_{d\text{-}c}t_1}{2C_1}$$

then substituting for $\frac{e_{i\,min}}{2}$ in the equation for $E_{d\text{-}c}$

$$E_{d\text{-}c} = \frac{e_{i\,max}}{2} + \frac{e_{i\,max}}{2} - \frac{I_{d\text{-}c}t_1}{2C_1} - I_{d\text{-}c}R_L$$

$$= e_{i\,max} - I_{d\text{-}c}\left(\frac{t_1}{2C_1} - R_L\right)$$

Since it is assumed that the capacitor charges nearly to the peak value, the conduction period of the valve must be small. Therefore $t_1 \doteqdot \frac{1}{2f}$ and $e_{i\,max}$ is nearly equal to E_{max}.

$$\therefore\ E_{d\text{-}c} = E_{max} - I_{d\text{-}c}\left(\frac{1}{4fC_1} - R_L\right).$$

3.14 Questions and problems.

1. Distinguish between L.T. and H.T. supplies to radio circuits.
2. Name the common primary sources of power for power supplies.
3. What voltage transformations are necessary in supplying the H.T. and L.T. to a transmitter from the d-c ships mains?
4. Define *rectification*.
5. Name three types of rectifiers.
6. Explain why a diode rectifies an a-c current.
7. Draw a simple half-wave rectifier with the accompanying waveforms of input voltage, output voltage and output current.
8. Define full-wave rectification.
9. Draw the circuit and explain the action of a full-wave rectifier.
10. Compare half-wave and full-wave rectification.
11. Why is it necessary to use a filter after rectification in receiver and transmitter power supplies?

12. Define *ripple* in the output of a rectifier.
13. Where is
    (a) a capacitor
    (b) an inductor placed in a rectifier circuit in order to reduce the ripple?
14. Describe the action of a simple capacitor filter
    (a) from a physical viewpoint
    (b) from a mathematical viewpoint.
15. Draw the waveforms of the output voltage and the current through the valve in a half-wave rectifier having a simple capacitor filter.
16. With a given filter capacitor what is the effect on the ripple in the output when the frequency of the a-c input is increased?
17. What are the effects of increasing the capacitance of the capacitor in a simple capacitor filter?
18. Define percentage ripple.
19. What is the percentage ripple in an output of $300V$ d-c if the peak ripple voltage is $15V$?
20. What are the advantages and disadvantages of using an inductor rather than a capacitor as a filter?
21. What fraction of the 120-cycle ripple fed to an L-section filter appears in the output if $L = 20H$ and $C = 8\mu F$?
22. Explain the action of a π-section filter from an impedance standpoint.
23. Explain the term *voltage regulation* as applied to a power supply.
24. Explain why a capacitor-input filter has poorer regulation than an inductor-input filter.
25. Under what conditions may each of the two types be used?
26. Why is a *load resistor* placed across the output terminals of a choke-input filter?
27. What is the advantage of using a *swinging choke* in a choke-input filter?
28. Give three reasons why a resistor may be connected across the output terminals of a power supply in parallel with the load.
29. Calculate the resistance and wattage rating of a resistor which is to act as a load resistor in a power supply which delivers $250mA$ to a load at $450V$.
30. If the above resistor is to be used as a voltage divider to supply $300V$ and $50V$ to loads which draw no current, calculate the resistance of each of the parts into which the whole resistor must be divided.
31. Recalculate the sizes of these parts if load currents of $75mA$ and $30mA$ are drawn from the $300V$ and $50V$ taps respectively.
32. Draw the circuit diagram of a receiver power supply giving suitable values for the size of components where possible.
33. What are the advantages and disadvantages of a bridge rectifier circuit over
    (a) a conventional full-wave rectifier
    (b) a half-wave rectifier?
34. Draw a bridge rectifier circuit using
    (a) valve symbols
    (b) metal rectifier symbols
    (c) an actual metal rectifier in which the four rectifiers are built in one unit.
35. What is meant by *peak inverse voltage* and what is its significance in connection with metal rectifiers?
36. Describe the construction of a copper oxide metal rectifier suitable for use in a power supply.
37. Compare the three types of metal rectifiers under the headings
    (a) advantages
    (b) disadvantages.

38. Where are mercury-vapour diodes commonly used and why?
39. What is meant by cathode disintegration and what causes it?
40. What precautions must be taken when using a mercury-vapour diode?
41. What is meant by *conditioning* a valve and when should this be done?
42. Draw the circuit diagram of a mercury-vapour-rectifier power supply.
43. (a) Draw the diagrams of a full-wave and a half-wave voltage doubler.
    (b) When is each type of circuit usually used?
    (c) Explain the operation of each circuit.
44. Why are voltage regulators necessary?
45. Name the different types of voltage regulators and list them according to the type of voltage changes which they can regulate.
46. Draw the circuits and explain the operation of each type of voltage regulator.
47. Distinguish between the two types of vibrators.
48. Why is a timing capacitor necessary? Explain its action.
49. Name and draw the waveforms for three of the faults which may occur in a vibrator.
50. Draw the circuit for an r-f oscillator power supply. Explain why it is safer than a conventional iron-core transformer type of rectifier power supply.

CHAPTER IV

THE TEST OSCILLOSCOPE

A description of the construction of the cathode-ray tube is given in Chapter I. This tube is widely used in electronic test equipment, radar displays and television. Cathode-ray tubes are divided into two types, electrostatic and electromagnetic depending on the method of deflection used. Electrostatic deflection is commonly employed in test oscilloscopes. The following section is devoted to a description of the auxiliary circuits which control the c.r.t. in a simple type of oscilloscope containing a three-inch cathode-ray tube.

4.1 C.r.t. control circuits.

A. The Electron Gun

The *electron beam* is produced by the *electron gun* which is located at the neck of the c.r.t. and which consists of the cathode, control grid, focusing anode and accelerating anode (Fig. 4-1). These electrodes receive operating potentials from a voltage divider R_1-R_5 connected across a d-c power supply. The supply voltage of approximately $600V$ is connected to ground at the positive terminal. All electrodes except the accelerating anode therefore receive negative potentials. In this way, the accelerating anode is operated at a high positive voltage with respect to the cathode, but is at ground potential to avoid modulation of the electron beam by electrostatic interference. The electron beam passes first through the control grid which is the most negative electrode in the tube. The number of electrons passing through the aperture in the control grid is determined by the grid-to-cathode potential. This potential can be varied by the *intensity* or *brilliance* control, R_5, and in this way the brightness of the spot of light on the screen can be regulated.

The sharpness of the spot of light on the screen can be varied by means of the *focus* control R_2. Since there is some interaction between the focus and intensity controls, R_2 and R_5 should be adjusted together for sharpest focus. Optimum focusing is most easily obtained when the beam current is small. The fixed resistors R_1, R_3 and R_4 are used to set the operating potential of the cathode and to limit the range of potentials over which R_2 and R_5 can operate.

B. Electron-Beam Positioning Circuits

The circuit shown in Fig. 4-2 is called an *unbalanced* beam-positioning circuit because one of the vertical-deflection plates and one of the horizontal-deflection plates are at ground potential. The

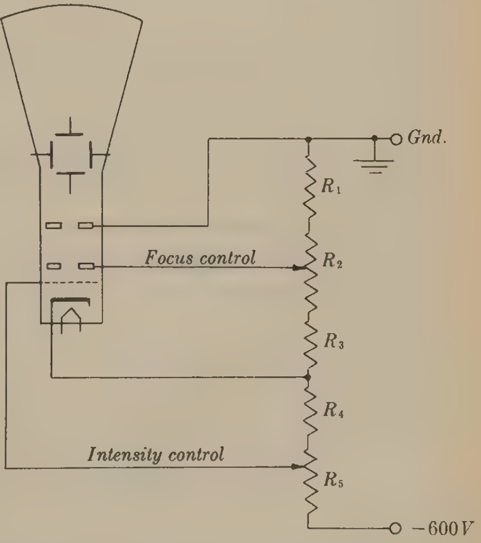


FIG.4-1 Electron beam control circuits.

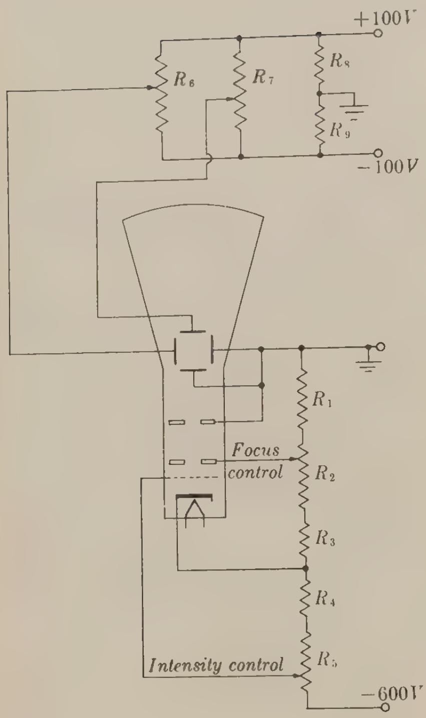


FIG. 4-2 Electron beam positioning circuits.

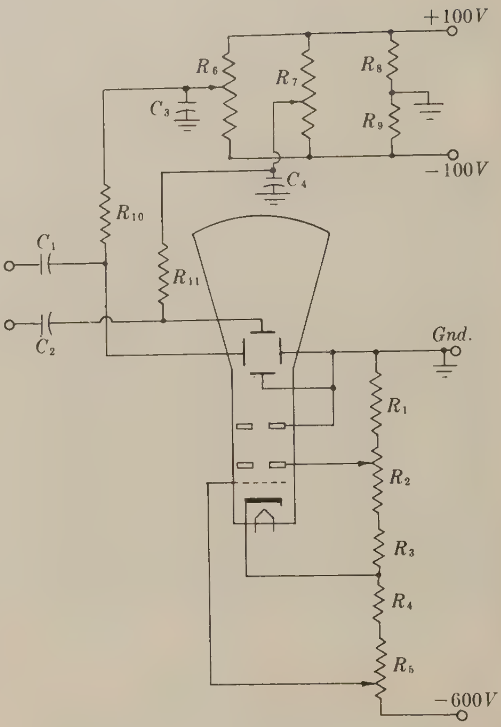


FIG. 4-3 Signal circuits.

centering voltages are applied to the other plate of each pair. As a result, the d-c voltages on the deflection plates are unequal.

The *horizontal-centering* control R_6 is connected to a source of approximately +100 volts at one end and −100 volts at the other end. Ground potential is midway between these extremes. When the variable tap of R_6 is set at the centre, the potential applied to the left X-plate is zero and the electron beam is not deflected to right or left of centre. If the variable tap of R_6 is moved upward a positive potential is applied to the left X-plate. This attracts the electron beam causing it to strike the screen to the left of centre. When the tap is moved below centre, a negative potential is applied to the left X-plate and the electron beam is deflected to the right of centre. By adjustment of R_6 the spot of light may be directed to any point along a horizontal line from the extreme left to the extreme right side of the c.r.t.

The *vertical-centering* control R_7 receives its potential from the same source as R_6. This control is used to apply a positive or negative potential to the upper Y-plate for vertical centering. By employing both the horizontal- and vertical-centering controls, the electron beam may be directed to any point on the fluorescent screen.

C. *A-C Signal-Voltage Circuits*

In addition to the d-c centering potentials, a-c voltages may be applied to one of each pair of deflection plates. *Horizontal deflection* of the beam is accomplished (Fig. 4-3) by applying an a-c signal through C_1 across R_{10} to the left X-plate. During the positive half of this cycle, the electron beam is attracted to the left of centre and during the negative half of the cycle it is repelled to the right of centre. The beam is therefore *swept* progressively from one side of the screen to the other. If the frequency of this *sweep voltage* is increased above about 10 cycles per second, the spot of light on the face of the c.r.t. begins to appear as a horizontal line. This is due partly to the *persistence of vision* of the human eye which cannot distinguish separate images that recur more frequently than about 30 times per second. The *persistency* of the fluorescent screen of the c.r.t. is also a factor in causing the rapidly moving spot of light to appear as a continuous line or *trace.*

Vertical deflection of the beam is accomplished in a similar way by applying an a-c signal through C_2 across R_{11} to the upper Y-plate. Such a signal causes a vertical trace to appear on the face of the tube when no horizontal-deflection signal is applied to C_1. C_3 and C_4 are by-pass capacitors which prevent a-c voltages from appearing across R_6 and R_7.

4.2 The sweep voltage. One of the most extensive uses of the cathode-ray oscilloscope is for the observation of voltage waveforms in electrical circuits. A graph of an a-c voltage waveform with voltage plotted vertically and time plotted horizontally may be produced on the screen of the tube. Such a graph or *oscillogram* is shown in Fig. 4-4.

To represent time horizontally on the c.r.t. screen, a sweep voltage is applied to the X-plates. This sweep voltage causes the electron beam to be deflected across the tube from left to right in a continuous uniform motion. When the electron beam reaches the right side of the screen it is made to *fly back* almost instantaneously to the left side of the screen to begin the next sweep. This motion is described as the *linear sweep* or *time base* of the cathode-ray tube. The waveform of the *sweep voltage* is illustrated in Fig. 4-5. This waveform is called a *sawtooth* voltage. The circuit which produces such a voltage is referred to as a *sawtooth generator* or *linear time base.*

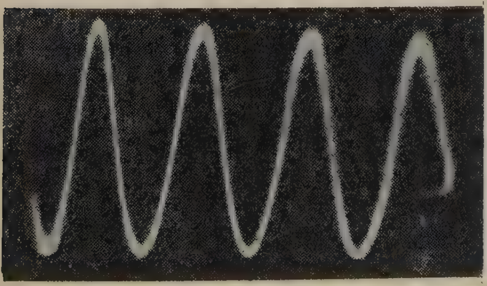


FIG.4-4 Oscillogram of an a-c voltage.

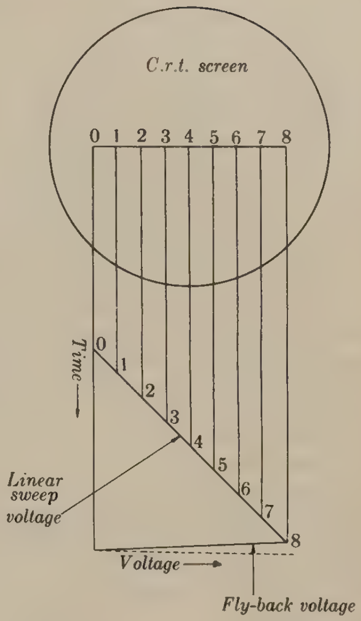


FIG.4-5 Oscillogram produced when a sawtooth sweep voltage is applied to the horizontal-deflection plates.

At time 0 the spot is at the left of the screen. The voltage applied to the deflection plates increases and this causes the spot to move across the screen. As the sweep voltage continues to increase at a uniform rate, the spot continues to move across the screen uniformly. When position 8 is reached the sweep voltage almost instantaneously is made to return to its initial value. Therefore the spot flies back to the left side of the screen and is ready to begin its uniform motion across the screen when the sweep voltage is re-applied.

It should be remembered that the electron stream strikes the screen at only one point at any instant. Many individual spots of light are produced on the screen in such rapid succession and so close together that the eye is incapable of distinguishing them separately. Moreover, the pattern is retraced on the screen so frequently that it appears as a continuous trace.

If a sinusoidal voltage is now applied to the *Y*-plates and the linear sawtooth voltage is applied to the *X*-plates the spot is deflected both vertically and horizontally at any instant. The combined effect is shown in Fig. 4-6 and a sinusoidal waveform is produced on the screen of the cathode-ray tube.

4.3 The simple sweep circuit. The simple *sweep circuit* (Fig. 4-7) used to develop a linear sawtooth voltage employs a thyratron valve. The variable resistor *R* is connected in series with the valve across the supply voltage. The sweep capacitor *C* is connected in parallel with the valve. The grid is biased negatively with respect to the cathode by the bias battery.

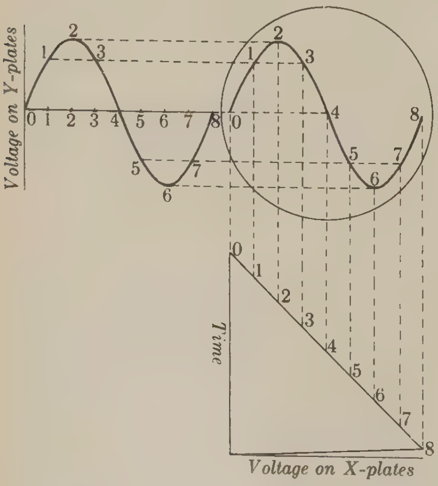


FIG. 4-6 Production of sine wave on c.r.t.

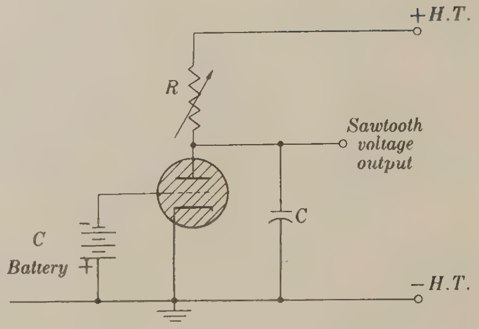


FIG. 4-7 Simple sweep circuit.

When the H.T. is applied electrons flow into the lower plate of the capacitor *C* making it negative. The electron flow from the upper plate of *C* through *R* causes a rising positive potential on this plate. The rate at which this voltage rises is determined by the time constant of *C* and *R*. The potential on the anode of the valve is the same as that on the charging capacitor. When this potential has risen to about 10 times the negative grid voltage, ionization occurs in the gas which fills the valve. The valve then conducts heavily and the sweep capacitor discharges rapidly through the valve. This causes the voltage across the capacitor and hence the anode potential to fall almost instantaneously. When the anode potential has fallen to the de-ionization potential, the valve stops conducting, the sweep capacitor begins to charge again and the cycle is repeated. The variation in voltage across the sweep capacitor is shown in the graph of Fig. 4-8. The charging period *ab* is part of an exponential curve rising toward the supply potential. Ionization is made to occur when the anode voltage has risen to about 10% of the total supply voltage. Over this part of the curve the voltage rises at a uniform rate for practical purposes. The de-ionization period is indicated by *bc* when the voltage across

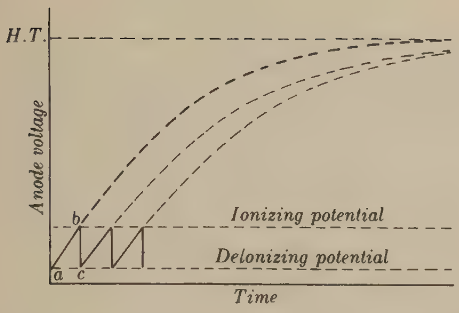


FIG.4-8 Graph showing variation in voltage across the sweep capacitor C of Fig. 4-7.

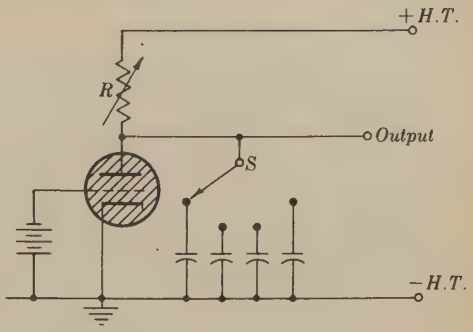


FIG.4-10 Sweep circuit with coarse and fine control of sweep frequency.

the sweep capacitor falls abruptly. This sawtooth waveform of voltage is applied to the grid of a resistance-capacitance coupled Class A amplifier. The output voltage (Fig. 4-9) is amplified and inverted. It is applied to the left X-plate of the c.r.t. to produce the linear sweep and *flyback*.

The time required to complete a sweep depends upon the rate of charge of the sweep capacitor. Therefore, the *sweep frequency* can be adjusted by making C or R variable. If either of these is increased the rate at which C charges is reduced. The sweep takes place more slowly and so the sweep frequency is reduced. To provide a range of sweep frequencies, it is common practice to include several capacitors of different sizes in the circuit. These are selected by the *sweep-selector* switch S which is the *coarse control* on the sweep frequency (Fig. 4-10). *Fine control* of the sweep frequency is then provided over each band by the variable resistor R.

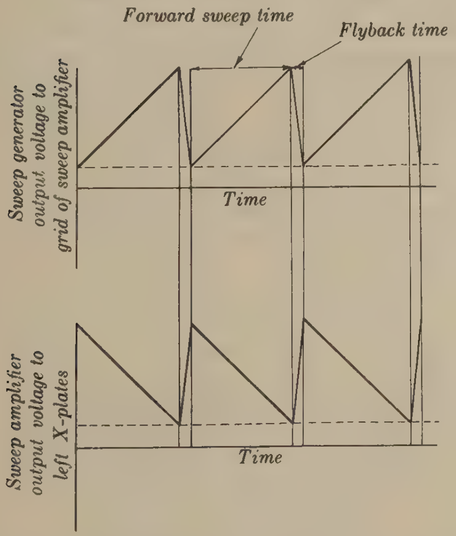


FIG.4-9 Output voltage of sweep circuit of Fig.4-7.

4.4 Synchronization of sweep circuits. If the pattern on the screen is to remain stationary the start of the sweep must always occur at the same time in each cycle of signal voltage. The period of the sweep must be exactly equal to the period of the signal or some whole multiple thereof. When this occurs the sweep is said to be *synchronized* with the input voltage. If the period of the sweep differs slightly from the period of the signal voltage, the pattern will drift across the screen of the oscilloscope, as illustrated in Fig. 4-11. The signal voltage shown is a sine wave of period slightly longer than the period of the sawtooth sweep. In the pattern shown in (a), only the part of the sine wave included between 0 and 1 will appear. On the second sweep the part between 1 and 2 will show, but since it differs from the part shown in (a), a slightly different picture will appear on the screen. In succeeding sweeps the picture will continue to change as shown in the

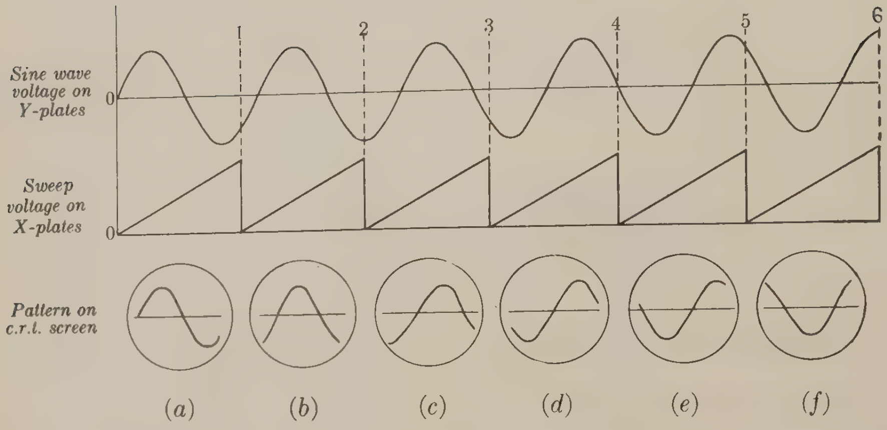


FIG.4-11 Drift of oscillogram pattern resulting from period of signal being greater than period of sweep.

diagram. The peak of the sine wave in each succeeding diagram is shifted somewhat to the right and as the sequence continues the pattern appears to travel across the screen from left to right. The speed at which the pattern appears to move depends upon the difference in frequency between the signal and the sweep.

Similarly if the period of the sawtooth sweep is longer than the period of the input signal the pattern appears to travel across the screen from right to left.

Sometimes it is desirable to observe more than one cycle of the signal on the screen. In this case the sweep must be synchronized at some frequency which is a sub-multiple of the signal frequency. If the sweep frequency is exactly one-half the signal frequency, two cycles of the sine wave will appear on the screen. If the flyback time is neglected, the number of complete cycles appearing on the screen is equal to the signal frequency divided by the sweep frequency.

The sweep frequency may be synchronized with the signal frequency by applying a small voltage of the signal frequency to the grid of the thyratron. The repetition rate of the sweep is then locked with the signal frequency. Synchronization depends on the fact that the anode voltage which produces ionization in the thyratron varies inversely with the voltage applied to the grid as illustrated in Fig. 4-12. Fig. 4-13 shows a thyratron sweep circuit with provision for synchronization.

The waveforms in Fig. 4-14 show the operating conditions of the thyratron and the effect of the synchronizing voltage on the sweep. With steady bias, ionization would take place at A. When part of the signal voltage is applied to the grid, the valve fires at a lower voltage B, because the grid has

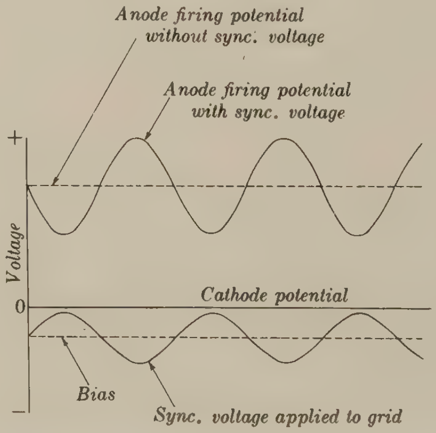

Anode firing potential without sync. voltage
Anode firing potential with sync. voltage
+
Voltage
0
Cathode potential
Bias
Sync. voltage applied to grid
–

FIG.4-12 Graph showing effect of varying grid voltage on anode firing potential.

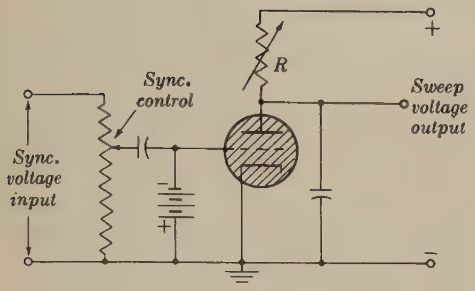


FIG.4-13 Thyratron sweep circuit with provision for synchronization.

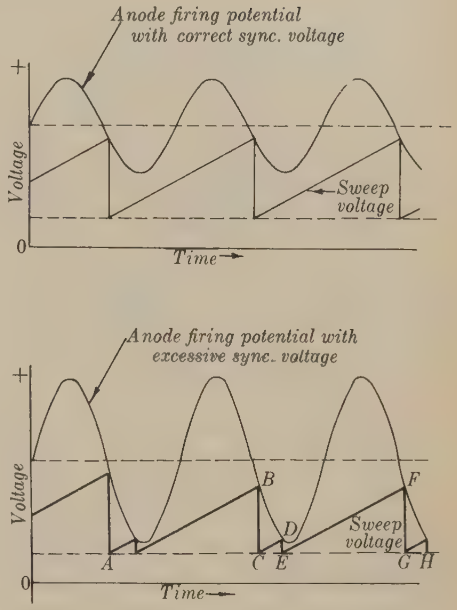


FIG.4-15 Distortion resulting from excessive synchronizing voltage.

been made less negative by the sync. voltage. The following sweep ends at *D* instead of *C* so that the period of the sweep is forced to be exactly the same duration as the period of the signal and a steady pattern will appear on the screen of the oscilloscope.

When using an oscilloscope the amount of synchronizing voltage employed should be a minimum. If the sync. control is turned too high, the pattern on the screen may be distorted. This is illustrated in Fig. 4-15. The use of a larger sync. voltage gives rise to larger variations in anode firing potential. The voltage across the valve rises from *A* to *B*. The valve fires at *B* and the sweep voltage drops to *C*. The voltage across the valve rises again and the valve fires at *D*. The voltage across the valve drops to *E* and starts to rise again. The sweep has been started at points *C* and *E* or twice in one cycle of the signal voltage. The pattern on the screen becomes disjointed and a continuous picture of the waveform cannot be obtained. In practice, the synchronization control should be turned to zero and the frequency of the sweep should be adjusted by the fine frequency control to achieve maximum stability of the pattern. If after adjustment, there is still a tendency for the pattern to drift across the screen, it may be stabilized by turning up the synchronization control just enough to stop the motion of the pattern.

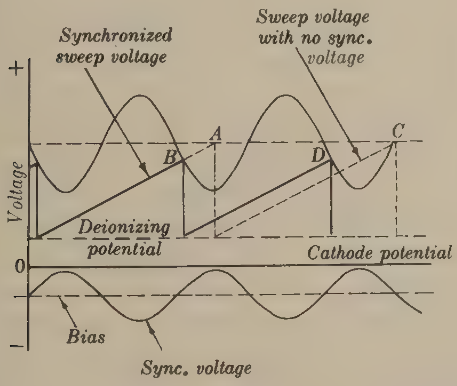


FIG.4-14 Operating conditions of thyratron with synchronizing voltage applied to grid.

4.5 A simple test oscilloscope. The block diagram of a simple oscilloscope for test purposes is shown in Fig. 4-16. A three-inch cathode-ray tube is used in an unbalanced circuit. One of each pair of deflection plates is connected to ground. The sweep generator produces a sawtooth output voltage which is amplified by the horizontal-deflection amplifier and applied to the left *X*-plate

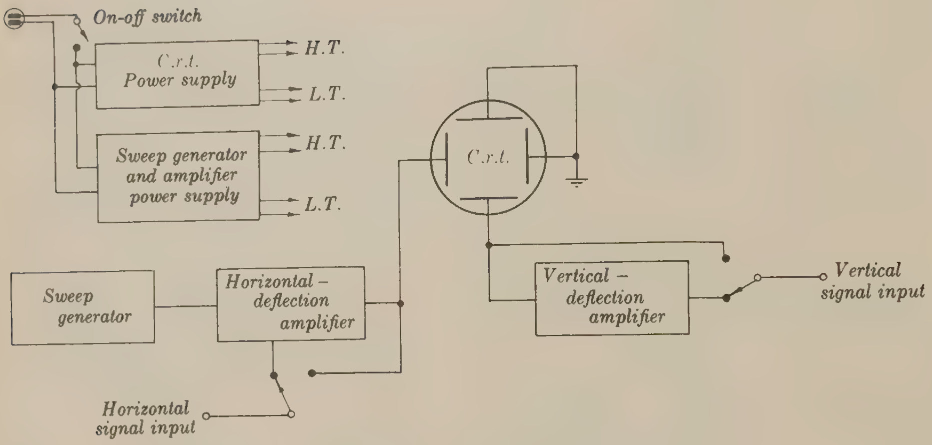


FIG.4-16 Block diagram of simple test oscilloscope.

of the c.r.t. Provision is also made for applying an external sweep voltage directly to the left X-plate or through the horizontal-deflection amplifier to this plate. The vertical input signal may be applied directly to the lower Y-plate or through the vertical-deflection amplifier to the Y-plate. A single transformer feeds two separate rectifier circuits. The high-voltage rectifier supplies H.T. to the cathode-ray tube circuits. The other rectifier supplies H.T. to the sweep generator and amplifiers.

A detailed circuit diagram of the oscilloscope is shown in Fig. 4-17 and a description of its operation follows.

A. Power Supplies

The oscilloscope is turned on by the on-off switch S_1 in the primary circuit of the power transformer T_1. This switch is incorporated in the intensity control. V_1 is a diode valve connected in a conventional half-wave rectifier circuit with its positive side connected to ground. A negative voltage of about 600 volts is developed across the filter capacitor C_{12} to supply the control voltages to the c.r.t. A separate secondary winding on the power transformer supplies V_2 in a standard full-wave rectifier circuit. The filter is a pi-section capacitor input filter consisting of C_{13}, C_{14} and L_1. Since the output is taken from the cathode of the valve, it is positive with respect to ground. This rectifier supplies about 300 volts to the sweep generator and amplifiers. Low-tension supplies for the heater of the c.r.t. and other valves are provided by two additional secondary windings on T_1.

B. Cathode-Ray Tube Controls

The negative voltage from the c.r.t. power supply is applied across the voltage divider R_{15}, R_{17}, R_{18}, R_{19}, R_{20}. The cathode is connected to a fixed voltage point between R_{19} and R_{20}. The grid is connected through R_{21} to the INTENSITY control R_{20}. The farther the variable tap is moved from the cathode end, the more negative the grid becomes with respect to the cathode and the fewer the electrons which are allowed to pass through the tube. Therefore R_{20} serves to control the brilliance or intensity of the trace on the screen. R_{18} is also a front panel control marked FOCUS. It is used to control the voltage on the first anode and hence the focus of the electron beam. The second anode is the accelerating anode. It operates

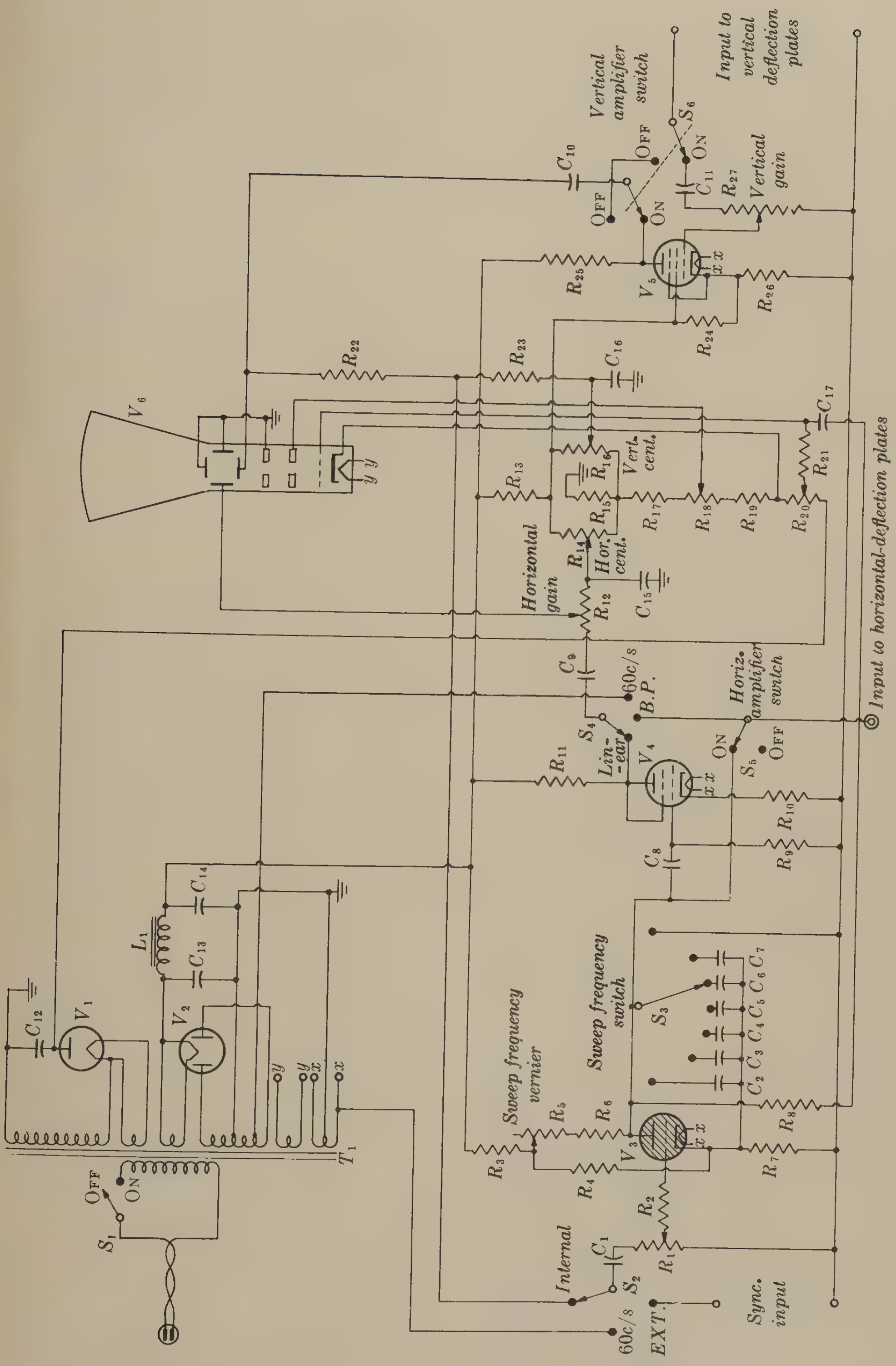


FIG. 4-17 Simple test oscilloscope circuit.

at ground potential and is several hundred volts positive with respect to the cathode and control grid. R_{14} is the HORIZONTAL CENTERING control. Its upper end is connected to a voltage divider R_{13}, R_{14}, R_{15} from +300 volts to ground and is therefore positive. The lower end is connected to the negative voltage divider $R_{15}-R_{20}$ and is therefore negative. R_{14} may therefore be used to shift the trace to the right or left of centre. The VERTICAL CENTERING control R_{16} operates in a similar manner to control the voltage on the lower deflection plate. This permits the trace to be shifted up or down on the screen.

C. Sweep Generator

A thyratron sweep circuit is used to provide a sawtooth waveform of voltage to the left plate of the c.r.t. The SWEEP FREQUENCY control S_3 selects the sweep frequency capacitor. The smaller the sweep capacitor, the more rapidly it charges and the faster the beam is swept across the screen. Fine control of the sweep frequency is provided by the SWEEP VERNIER control R_5. It provides a continuous variation of the rate at which the sweep capacitor charges over each range selected by S_3. R_2 is a limiting resistor to limit the flow of grid current in the thyratron. R_7 is the cathode bias resistor, part of the voltage divider R_3, R_4, R_7 across the H.T. supply.

A synchronizing voltage may be applied to the grid of the thyratron from any one of three sources selected by S_2. In the first position the sync. comes from the binding post marked EXTERNAL SYNC. In the second position the sync. is a sine wave of voltage from one side of the 6.3-volt filament winding on the power transformer. This locks the sweep in with the line frequency, usually 60 cycles per second. In the position marked INTERNAL the sweep frequency is made equal to, or a sub-multiple of, the signal frequency. This keeps the pattern from drifting across the screen. The sync. voltage is obtained from the voltage divider R_{22}, R_{23} across the input to the lower deflection plate. R_1 is the SYNCHRONIZING control on the front panel. It is adjusted so that the pattern remains stationary.

D. Horizontal-Deflection Amplifier

The sweep amplifier employs a tetrode valve connected as a triode in a Class *A* amplifier circuit. The sweep voltage from the thyratron is applied through C_8 across R_9 to the grid of V_4. The output is amplified and inverted, Fig. 4-18 (a) and (b). It is applied through C_9 to the left deflection plate of the c.r.t. The cathode resistor R_{10} is not bypassed. This introduces a measure of degeneration which maintains a fairly linear output voltage over a wide range of input signal amplitude. The sweep selection switch S_4 connects the output of the sweep amplifier to the deflection plates when in the position marked LINEAR. When switched to B.P. an external sweep may be applied from the external binding posts to the left *X*-plate. If S_4 is switched to 60 c/s a sine wave at the line frequency is applied from the power transformer through C_9 to the sweep plate of the c.r.t. The sweep is then not linear, but moves sinusoidally from left to right and returns at the frequency of the power input.

The horizontal-amplifier ON-OFF switch is ganged with the sweep frequency switch. If it is desired to amplify the external sweep voltage from the binding post, the sweep amplifier switch is put to the ON position and the sweep selector switch S_4 is set to LINEAR.

The length of the trace may be adjusted by means of the HORIZONTAL GAIN control. It allows the whole or any part of the voltage from the sweep amplifier to be applied to the left deflection plate. The by-pass capacitor C_{15} prevents the sweep voltage from modulating the voltage across R_{15}–R_{20} and interfering with the normal operation of the vertical centering, focus and intensity controls.

E. Vertical-Deflection Amplifier

The amplifier for the vertical deflection signals to the Y-plates is a Class A amplifier using a pentode valve V_5; the input from the binding post is applied through C_{11} and the VERTICAL GAIN control R_{27} to the grid of the valve. The output from the anode is applied through the ON-OFF switch in the ON position and through C_{10} to the lower Y-plate. When the vertical amplifier switch is turned to the OFF position the signal is applied through C_{10} without amplification to the deflection plate. The cathode is biased down by the voltage divider R_{13}, R_{24} and R_{26} across H.T. The screen also receives its potential from this voltage divider. R_{25} is the anode load resistor for this valve. C_{16} is a by-pass capacitor which by-passes vertical deflection signals from R_{16}.

R_8 and C_{17} from the anode of the thyratron V_3 to the grid of the c.r.t. provide a means of making the retrace or fly-back invisible on the c.r.t. This circuit operates at the end of the sweep when the thyratron is fired to provide the fly-back. The sudden discharge of the sweep capacitor through the valve causes the anode voltage to fall abruptly. This negative voltage is coupled through R_8, C_{17}, across R_{21} to the grid of the c.r.t. It cuts off the electron stream in the c.r.t. during the retrace time. C_{17} has a very small capacitance so that the time constant of R_8, C_{17} and R_{21} is small. This allows C_{17} to charge gradually during the forward sweep without causing any appreciable variation in the intensity of the beam. Fig. 4-18 (c) shows the waveform of the voltage which blanks out the retrace.

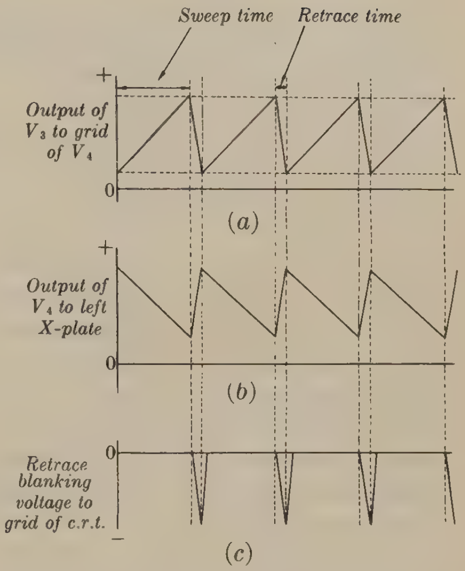

Sweep time
Retrace time
+
0
Output of V_3 to grid of V_4
(a)
+
0
Output of V_4 to left X-plate
(b)
0
−
Retrace blanking voltage to grid of c.r.t.
(c)

FIG.4-18 Output waveforms of V_3 and V_4 in Fig. 4-17.

4.6 Double-beam oscilloscopes. Single-beam oscilloscopes are used for the study of a single waveform at one time. However, two waveforms can be displayed on the tube using a single-beam oscilloscope if an electronic switch is put into the circuit. The waveforms to be compared are applied to the switch which then alternately applies sections of the waves to the single-beam oscilloscope. The eye sees two waveforms on the tube face.

However, this is an inconvenient arrangement since two units are required. Double-beam oscilloscopes have been developed to achieve the same result. Separate waveforms are applied to the double-beam oscilloscope. Each passes through a separate amplifier channel and is displayed separately on the screen. The time base of the two waves is the same so that they may be compared.

The double beam is obtained by placing a screen between the Y-plates to split the focused beam as it leaves the final anode, the splitter plate itself being directly connected to the final anode within the tube. Each Y-plate thus affects one half of the beam independently to provide the two vertical traces. Because of this construction, the two voltages, if they are electrically in phase, deflect the beams to display the waveforms as 180 degrees out of phase on the scope. To have the waveforms displayed in the correct relationship, a phase inverter is inserted in the amplifier circuit feeding one of the plates. The X-deflection plates provide the usual horizontal deflection and are common to both beams. The double-beam oscilloscope makes it possible to compare two waveforms simultaneously without phase delay or frequency limitations. The advantage of the double-beam oscilloscope becomes evident in making frequency comparisons, phase shift measurements and distortion analyses.

If it is required to compare frequencies which are relatively close, for example, the output from two oscillators one of which is under test, then a double-beam tube can be used most effectively. A stationary wave from the standard oscillator is displayed upon the screen and the second wave from the oscillator under test is superimposed upon it. Differences in frequency are indicated by the second wave moving horizontally with respect to the first. The period of the difference in frequency may be measured with a stop watch.

The phase difference between two waveforms of equal frequency may be measured by superimposing the waves on the double-beam tube. A graticule (scale) placed in front of the screen measures the phase difference by measuring the distance between corresponding peaks of the two waves and taking that as a fraction of the total wavelength.

Distortion that is put into a voltage wave by an amplifier can be observed by displaying the input and the output waveforms on a double-beam tube. Superposition of the waves for comparison is assisted by having amplifiers in each of the two plate circuits. Care must be taken that distortion is not introduced by the amplifiers of the oscilloscope.

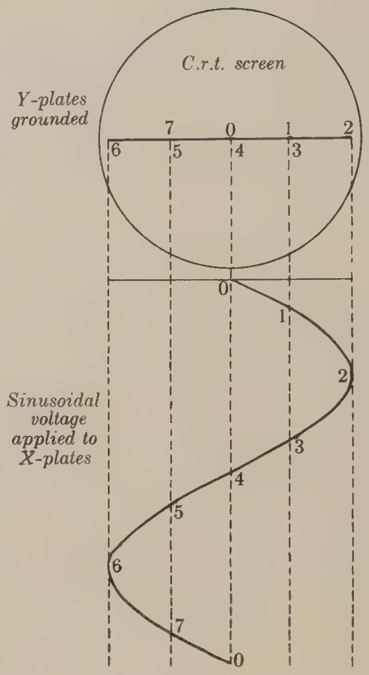


FIG.4-19 Oscillogram produced when a sinusoidal alternating voltage is applied to X-plates.

4.7 Lissajous figures.

A. Fundamental Patterns

A *Lissajous figure* is a pattern traced out on an oscilloscope screen when a-c voltages are applied to both the horizontal- and vertical-deflection plates. Examination of the pattern which appears on the screen provides a very satisfactory method for determining the frequency and phase relationships between the two voltages.

The simplest type of Lissajous figure is produced when the frequency ratio is 1:1, that is when the frequency of the signal applied to the vertical plates is the same as the frequency of the signal applied to the horizontal-deflection plates.

If a sinusoidal alternating voltage (Fig. 4-19) is applied to the *X*-plates while the *Y*-plates are grounded, a single horizontal line appears on the screen. At zero voltage the spot of light is at the centre of the screen. A moment later marked 1, the spot has moved to the right. After one-quarter of a cycle, it is at the right-hand side of the screen, 2. Then as the voltage decreases, the spot of light retraces this path to the centre point. During the negative half cycle the spot moves from 4 to 6 and back to 0 again. If the frequency of the alternating voltage is sufficiently high, the pattern appears as a continuous horizontal straight line.

Similarly, if a sinusoidal alternating voltage is applied to the vertical-deflection plates while the horizontal-deflection plates are grounded, a single vertical line appears on the screen.

If alternating voltages of the same frequency and amplitude and in the same phase are applied to the horizontal- and vertical-deflection plates a single oblique line making an angle of 45° with the *X*- and *Y*-axes appears on the screen (Fig. 4-20).

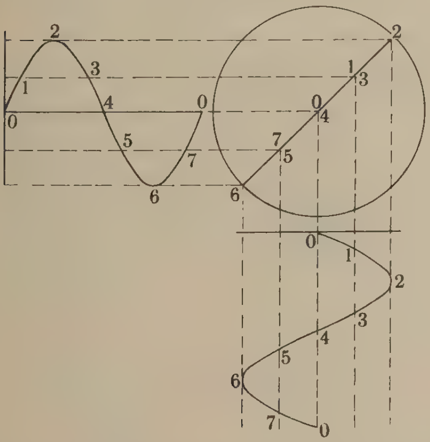


FIG.4-20 Oscillogram produced when sinusoidal alternating voltages of same frequency, amplitude and phase are applied to *X*- and *Y*-plates.

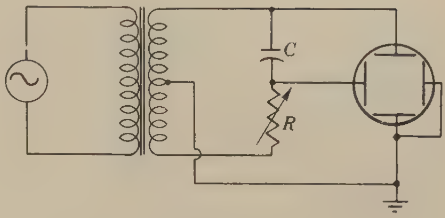


FIG.4-21 Phase-splitting circuit.

When the voltages applied to both *X* and *Y* plates are zero, the spot is at the centre of the screen. A moment later, marked 1, the spot has been moved upward by the voltage on the *Y*-plates and to the right by the voltage on the *X*-plates. After one-quarter of a cycle it is at the upper right-hand corner of the screen marked 2 and during the remainder of the cycle moves through points 3, 4, 5, 6, 7, 0 in succession. Thus the combination of two equal voltages in the same phase and of the same frequency results in a straight line making an angle of 45° with the *X*- and *Y*-axes.

If the two voltages are of the same frequency and phase but of unequal amplitudes a straight line appears at some angle other than 45°. The line is tilted toward the axis to which the greater voltage is applied.

B. Phase Determination

By means of the *phase-splitting* circuit shown in Fig. 4-21, phase shifts from 15° to 170° can be obtained. If *R* is adjusted so that there is a phase difference of 90° between the sine wave voltage applied to the vertical- and horizontal-deflection plates the pattern on the screen is a circle. This is illustrated in Fig. 4-22.

At 0 the horizontal deflection voltage is zero but the vertical deflection voltage is a maximum and the spot of light appears at the top of the screen. During the next quarter cycle the horizontal deflection voltage increases to a

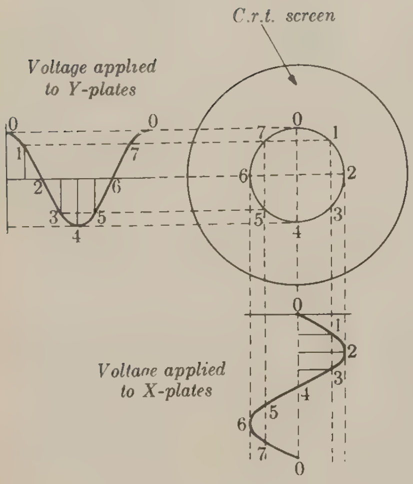


FIG.4-22 Oscillogram produced when sinusoidal alternating voltages of same frequency and amplitude but differing in phase by 90° are applied to X- and Y-plates.

maximum and the vertical deflection voltage decreases to zero. The spot of light moves successively through points 1, 2, 3, 4, 5, 6, 7, 0, to trace out a circle on the screen. If the amplitudes of the voltages applied to the vertical- and horizontal-deflection plates are unequal an ellipse appears instead of a circle. The circuit shown in Fig. 4-21 is usually modified as shown in Fig. 4-23 to include potentiometer R_1. This permits adjustment of the relative amplitudes of vertical and horizontal deflection voltages.

If sine wave voltages of the same frequency and amplitude are applied to vertical- and horizontal-deflection plates and the phase difference between them is varied from 0° to 360° a series of Lissajous figures similar to those shown in Fig. 4-24 is produced.

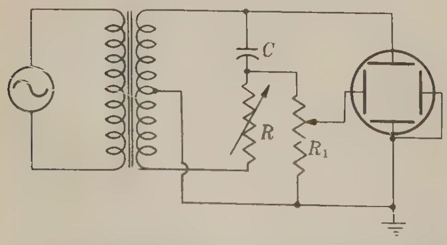


FIG.4-23 Phase-splitting circuit with provision for varying the relative amplitudes of voltages.

When the voltages are exactly in phase, Fig. 4-24 (a), the trace is a straight line which makes an angle of 45° with the axes. As the phase difference is increased, the straight line opens into a broadening ellipse. When the phase difference is 90°, Fig. 4-24 (d),

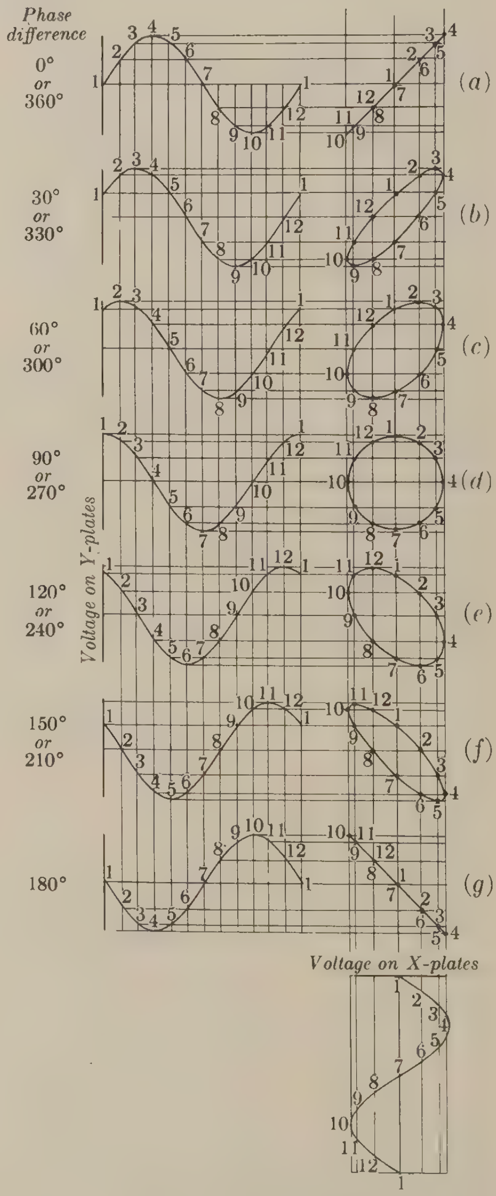


FIG.4-24 Lissajous figures.

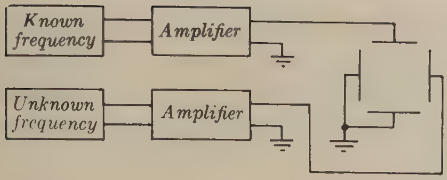


FIG. 4-25 Block diagram of arrangement for determining complex frequency ratios.

the ellipse becomes a circle. As the phase difference is increased from 90° to 180° the circle becomes an ellipse and the ellipse closes to a straight line but the line makes an angle of 135° with the positive direction of the X-axis. Observation of the Lissajous figures produced by two voltages provides a simple method for determining the phase angle between them.

C. *Frequency Determination*

1. Completed-Loop Method.

Lissajous figures provide an excellent means for determining the frequency ratio of two alternating currents. It is the simplest method for determining complex frequency ratios. A block diagram of the basic circuit used for this purpose is shown in Fig. 4-25. Fig. 4-26 represents a very simple example of the pattern produced by two sine wave voltages which differ in frequency. The voltages applied to the X- and Y-plates are of equal amplitude, in phase with each other at time 0 but differ in frequency in the ratio of 1:2. A figure 8 pattern is obtained.

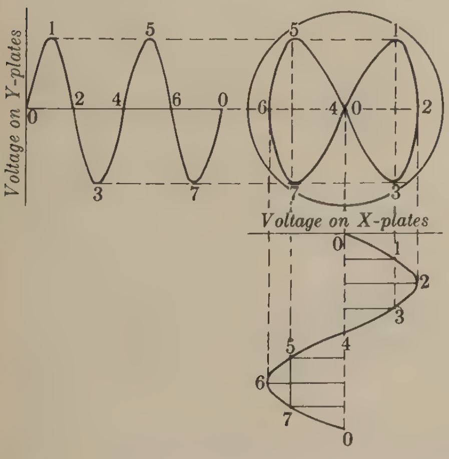


FIG. 4-26 Oscillogram produced by two sine wave voltages differing in frequency.

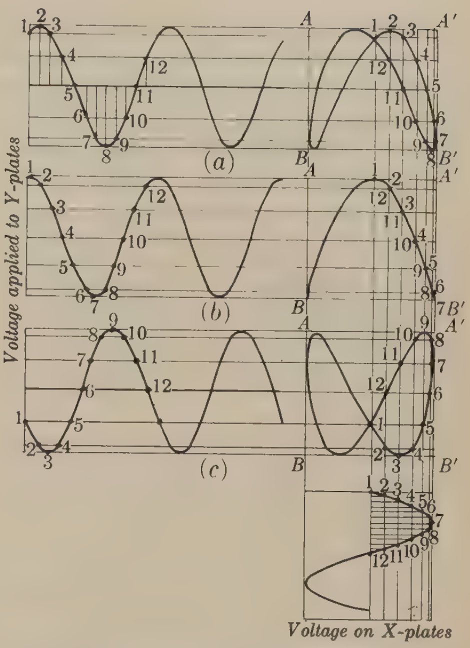


FIG. 4-27 Variation in oscillogram pattern with changes in phase.

In examining Lissajous figures it is helpful to consider them to be a side view of a pattern traced on a transparent cylinder on which the observer may view the whole wave at all times as it travels around the cylinder. The pattern on the cylinder may appear to rotate as a result of changes in the relative phases of the two voltages. Fig. 4-27 illustrates some of the patterns obtained for a frequency ratio of 1:2. The actual frequencies are not important as long as the frequency of the voltage applied to the Y-plates is twice that applied to the X-plates. If the two voltages are not in phase the pattern changes as shown in Fig. 4-27 (a), (b) and (c). When the phase difference is 90° the pattern closes as in (b). If the phase difference is greater than 180° the patterns are inverted as in (c).

The common feature of all these patterns which provides a method for determining the frequency of an unknown signal is that the vertical line AB or $A'B'$ is tangent to the pattern at one point while the horizontal line AA' or BB' is tangent to the pattern at two points indicating the ratio 1:2. This is true even for the line AA' in (b) since the trace passes the point of tangency twice during each cycle. The rule for frequency determination may be stated as follows:

$$\frac{\text{Frequency of voltage applied to } X\text{-plates}}{\text{Frequency of voltage applied to } Y\text{-plates}} = \frac{\text{No. of points at which the pattern is tangent to a vertical line}}{\text{No. of points at which the pattern is tangent to a horizontal line.}}$$

Thus, in frequency determination tests, a voltage of known frequency is applied to the X-plates while the unknown voltage is applied to the Y-plates. The Lissajous figure is observed and the unknown frequency is determined by substitution in the above formula. For example, in Fig. 4-28 (d) the pattern touches the vertical line at two points and the horizontal line at three points.

Hence, $$\frac{f_x}{f_y} = \frac{2}{3}.$$

If the known frequency applied to the X-plates is 500 c/s then the unknown frequency (f_y) is obtained by solving the equation

$$\frac{500}{f_y} = \frac{2}{3}$$

$$f_y = 750$$

Therefore the unknown frequency is 750 c/s.

In a similar manner, the ratio of the applied frequencies for each of the patterns indicated in Fig. 4-28 may be determined. When one frequency is not an integral multiple of the other, the successive cycles of the lower frequency do not fall upon each other in the pattern. If the trace of Fig. 4-28 (i) is followed, it is necessary to go across the screen and back again twice before the trace

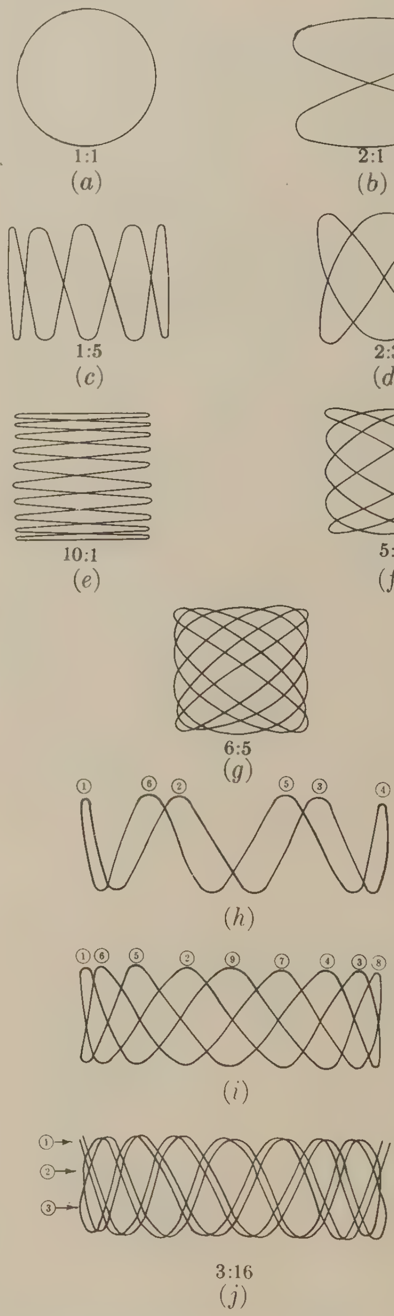


Fig. 4-28 Completed-loop method of frequency determination.

begins to repeat itself. This is called a *two-line* pattern and has a frequency ratio of 2:9. The vertical frequency is therefore $4\frac{1}{2}$ times the horizontal frequency. The same analysis shows that in Fig. 4-28 (j) three cycles of the horizontal frequency are completed before the pattern begins to repeat and this is therefore a three-line pattern with a frequency ratio of 3:16.

When the pattern is very complex it becomes difficult to count the number of peaks and end loops on the screen due to the apparent rotation of the pattern. This handicap may be overcome by photographing the image on the screen. Since the usual green image does not photograph well, special cathode-ray tubes are available which have a chemical coating on the screen that produces a blue image. The blue has a much higher actinic value and the length of time required for proper film exposure is reduced considerably.

2. UNCOMPLETED-LOOP METHOD.

This method uses Lissajous figures which are produced when sawtooth voltage waveforms are applied to the *X*-plates and sine wave voltages to the *Y*-plates. When the frequency of the voltage applied to the *Y*-plates is a simple multiple of the sweep frequency the frequency ratio is easily determined as shown in Fig. 4-29. Fig. 4-29 (f) illustrates the pattern produced when the frequency ratio is 2:3. At the end of the first sweep one and one-half cycles have been traced on the screen. At this instant the spot of light jumps back to the left-hand side of the screen. During the next sweep the succeeding one and one-half cycles are traced out to form the pattern illustrated.

Fig. 4-30 represents the patterns produced when the frequency ratios are larger. The frequency ratio can be determined from the following formula:

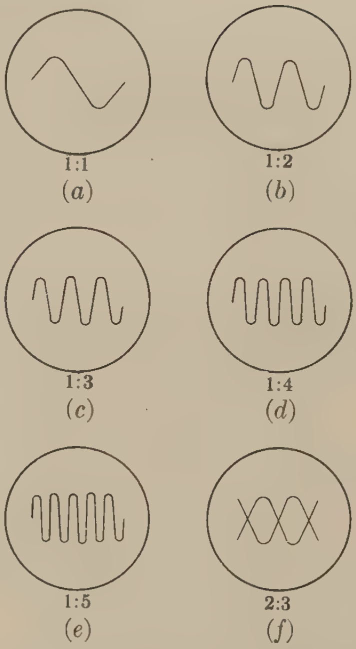


FIG. 4-29 Uncompleted-loop method of frequency determination.

$$\frac{\text{Frequency of sweep}}{\text{Frequency of voltage applied to } Y\text{-plates}} = \frac{\text{No. of times line } AB \text{ is intersected}}{\text{No. of points at which the pattern is tangent to a horizontal line.}}$$

These patterns are convenient for calibrating a linear sweep but in general the completed-loop pattern is preferable to the uncompleted-loop pattern for frequency determination.

4.8 Uses of the cathode-ray oscilloscope. Several applications of the oscilloscope in monitoring and testing circuits are discussed in the preceding sections. The visual observation of voltage and current waveforms is a valuable aid in understanding the operation of oscillators, amplifiers, rectifiers and modulators. The oscilloscope permits the accurate measurement of frequency. The vertical pattern produced by the unknown frequency is in effect compared with the known horizontal sweep frequency. The oscilloscope is also useful in determining

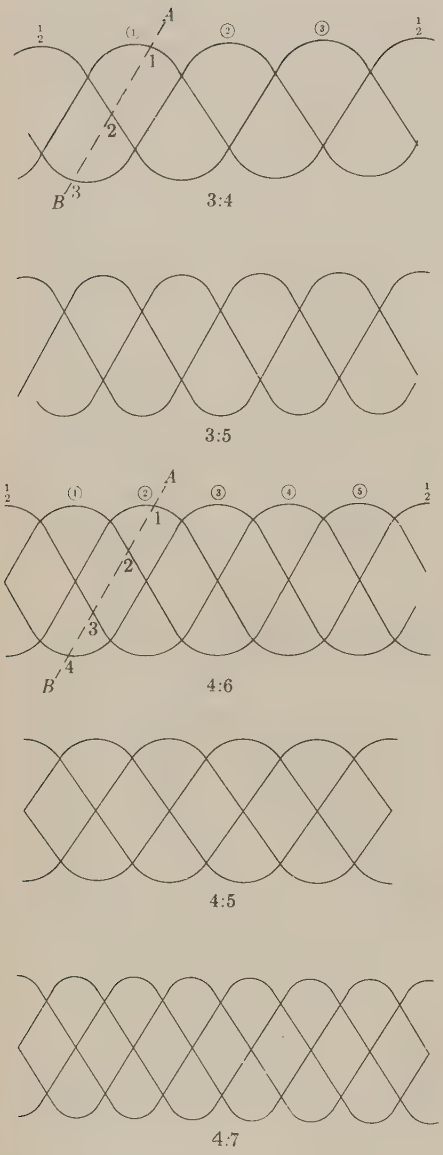


Fig. 4-30 Uncompleted-loop method of frequency determination when frequency ratios are large.

the phase relationships between voltage and current in alternating current circuits. Other uses of the oscilloscope are described in the following paragraphs and in subsequent chapters.

The cathode-ray tubes that employ electrostatic deflection are voltage-operated instruments. The amount of deflection of the electron beam on the screen is proportional to the deflecting voltage. Hence the tube may be used as a voltmeter for either a-c or d-c measurements. However, since both the range of voltage and the accuracy of indication are less than that available in commercial d-c voltmeters, the oscilloscope is not widely used for measurement of d-c voltages. It is much more useful for measuring alternating voltages.

The oscilloscope is first calibrated by applying a known voltage to the vertical-deflection plates and observing the amount of deflection of the beam from its zero position. The sensitivity of cathode-ray tubes varies widely according to size of tube and accelerating voltage used. It ranges roughly from 5 to 100 millimetres deflection per 100 volts applied. If the voltage to be measured is applied through an amplifier the effective sensitivity is considerably increased. Readings must of course be taken with the gain control at the same setting as that at which the instrument was calibrated.

A-c voltmeters of the current-operated type measure the effective value of the a-c voltage. If the peak voltage is required, the effective value may be multiplied by 1.414 provided that the waveform is sinusoidal. Otherwise, a vacuum-tube voltmeter should be used which measures the peak voltage. Its scale is usually calibrated to give the effective value of the a-c voltage assuming that the waveform is sinusoidal. The oscilloscope however presents the waveform visually and permits easy determination of the peak value of the voltage regardless of waveform. The input impedance of the oscilloscope is high and the frequency range is large in a well designed instrument.

The oscilloscope may be used as a current-measuring device indirectly. The current to be measured is passed through a resistance of known value and the voltage across the resistance presented on the screen. The current can be determined by substitution in Ohm's Law, $E = IR$.

The power consumed in a resistive load may also be determined by measuring the current and voltage and substituting in the power formula, $P = IV$.

4.9 Questions and problems.

1. Draw the diagram of an unbalanced beam-positioning circuit connected to a c.r.t.
2. What is the chief disadvantage of the unbalanced circuit?
3. How is time represented horizontally on a c.r.t. screen?
4. What purpose is served by a linear sweep on the horizontal-deflection plates? Why is it called a linear sweep?
5. By means of a diagram explain the pattern on a c.r.t. screen when a linear sweep is applied to the X-plates and a sine wave voltage is applied to the Y-plates.
6. Draw the circuit diagram of a simple sweep circuit and explain its operation.
7. What is meant by synchronization of the sweep circuit? Why is it needed?
8. Draw the circuit of a sweep generator which includes a sync. voltage input.
9. With the aid of graphs explain how a sweep circuit is synchronized.
10. Explain how distortion of the pattern is produced by applying a sync. voltage which is too large.
11. Draw a block diagram of the circuit of a simple test oscilloscope.
12. (a) What pattern is obtained on the c.r.t. screen when two sine wave voltages of equal amplitude and in the same phase are applied, one to the X-plates and one to the Y-plates of the c.r.t.?
    (b) What is the result when the amplitudes of the two voltages becomes unequal?
    (c) What is the result of changing the phase relationship of the two equal sine wave voltages?
13. What is the rule for frequency determination by means of Lissajous figures using
    (a) the completed-loop method
    (b) the uncompleted-loop method?
14. How may the oscilloscope be used as a voltmeter?

CHAPTER V

OSCILLATORS

An *oscillator* is a valve circuit which produces an alternating voltage output from a direct-current source. The frequency of oscillators varies over wide limits from a few cycles per second to several thousand megacycles per second. Audio-frequency oscillators are used in laboratory and test equipment, code practice sets and in modulators for m-c-w transmitters. Radio-frequency oscillators are used primarily in radio and radar receivers and transmitters.

The oscillatory circuit is described in Chapter XII, Vol. 1. If a charged capacitor is connected to the terminals of an inductor, energy circulates from one component to the other until it has been dissipated in heat losses, capacitor losses, induced eddy currents and radiation losses. The current that flows in such a circuit is a *damped* oscillatory current. If it is desired to maintain the oscillations, energy must be supplied to the oscillatory circuit at the correct frequency and in the proper phase to make up for the circuit losses. A valve is used to control the flow of energy from the direct-current source to the alternating-current circuit.

Several types of oscillators are described in the following sections. They are classified according to the position of the oscillatory circuit which may be placed

(i) between grid and cathode
(ii) between anode and cathode
(iii) between grid and anode.

5.1 The tuned-grid oscillator. The *tuned-grid oscillator* shown in Fig. 5-1 consists of an oscillatory circuit L_1C_1 connected between grid and cathode of a valve. A *reaction* or *tickler* coil L_2, connected in the anode circuit is inductively coupled to L_1, the two forming a transformer. Automatic bias is provided by the grid-leak resistor R and capacitor C_2.

A. Establishment of Oscillations

When the switch S is closed, current through the valve increases from zero. This current flowing through L_2 sets up an increasing magnetic field about L_2 which links the turns of L_1. Consequently a voltage is induced in L_1 which charges capacitor C_1. Energy is therefore supplied from the anode circuit to the grid circuit causing oscillations to begin in L_1C_1. The oscillating current set up in L_1C_1 develops an alternating voltage across C_1 which is applied between grid and cathode to control the flow of current through the valve.

The polarity of the grid voltage relative to the anode voltage is of prime importance in building up oscillations in the circuit. The windings of L_1 and L_2

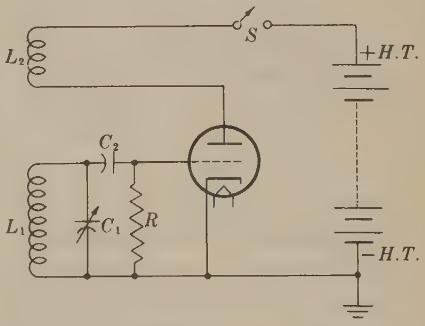


FIG.5-1 The tuned-grid oscillator.

are connected into the circuit so that these two voltages are in opposite phase. When current increases through the valve and L_2, causing the anode potential to fall, the voltage induced in L_1 causes the grid potential to rise. The rising grid potential permits an increased current to flow through the valve. This constitutes *positive feedback* or *regeneration.* That is, the cathode-to-anode current, which at first was limited by zero grid potential, is now increased by the positive grid potential.

The degree to which regeneration proceeds in this direction is limited by several factors. When the grid goes positive with respect to the cathode, current flows up through R to charge C_2. However, grid current also flows to charge C_2. Hence the current through R is limited and the grid potential is also limited to a small positive value with respect to the cathode. Therefore the anode current stops increasing. The positive grid potential may also result in anode current saturation which causes the anode current to stop increasing. The oscillating current in the tuned circuit now reverses, the p.d. across the capacitor C_1 decreases and the grid potential falls. The current through the valve therefore decreases. The anode potential begins to rise and a voltage is induced in L_1 which is of the same polarity as that produced by the oscillating current in L_1. Thus, regeneration takes place during the negative as well as the positive half of the alternating cycle in L_1C_1. Regeneration in this direction may continue until the valve is cut off. The grid voltage goes more negative until the oscillation reverses in the tuned circuit. The valve acts as an amplifier to increase the amplitude of the oscillations. A small amount of energy fed back from L_2 to L_1 controls a larger amount of energy flowing through the valve.

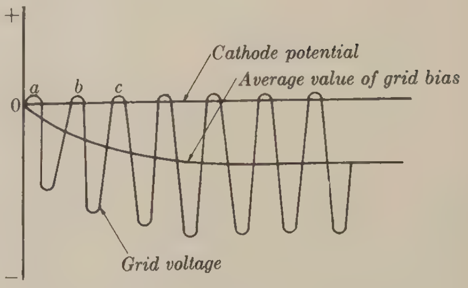


Fig.5-2 The establishment of grid bias.

B. Maintenance of Oscillations

The amplitude of oscillations in the tuned grid circuit continues to increase until limited by two factors, losses from the circuit and grid bias. The increase in bias is shown by the diagram of Fig. 5-2. During the first half cycle of oscillation indicated at *a*, the grid is made slightly positive to the cathode. Electrons flow from the cathode to the grid and accumulate on the right plate of C_2. During the time between *a* and *b* that the grid potential is below the cathode potential, C_2 discharges slightly through R causing a negative potential to be applied to the grid of the valve. The time constant of R and C_2 is made great enough that C_2 can lose only a small part of its charge in one cycle. When the grid potential again rises above the cathode potential at *b*, more grid current flows and the bias voltage developed across R is increased for the next cycle. As the amplitude of oscillations increases in the circuit the average value of grid bias increases. This bias limits the time during which current flows through the valve and so limits the amount of energy fed back from L_2 to L_1. In addition, the circuit loses energy each cycle due to radiation and heating effects. These losses are more pronounced as the oscillating current in the circuit increases. A point must therefore be reached at which losses, grid bias, and the restricted positive grid swing, prevent the amplitude of oscillations from increasing further.

The grid bias becomes steady when C_2 receives as much charge when the grid is positive as it loses through R during the rest of the cycle. The circuit continues to oscillate at a constant amplitude as long as operating conditions remain the same.

Automatic bias has the advantage over other types of bias in that it makes the oscillator self-starting. When the circuit is initially switched on, the grid bias is zero and current is permitted to flow. If operating conditions change while the circuit is oscillating, the bias increases or decreases in such a way as to maintain the oscillation.

The frequency at which the circuit oscillates is given for practical purposes by the formula

$$f = \frac{1}{2\pi\sqrt{LC}}$$

where f = frequency in cycles per second
L = inductance of L_1 in henries
C = capacitance of C_1 in farads.

Certain other factors such as the value of the supply voltage, valve factors, heating and loading affect the frequency to a lesser degree.

Since current and voltage amplitudes in the grid circuit are relatively small, low output power is available in the oscillatory circuit. This oscillator is suitable as a *heterodyne oscillator* in a receiver. By placing the oscillatory circuit in the anode circuit of the valve, larger power output becomes available.

5.2 The tuned-anode oscillator. Fig. 5-3 illustrates a *tuned-anode oscillator* in which the oscillatory circuit L_1C_1 is connected in the anode circuit of the valve. The tickler coil L_2 which is inductively coupled to L_1 is connected between the grid and cathode. When the switch is closed in the H.T. circuit, the current through the valve increases from zero. This sets up an increasing magnetic field around L_1 and charges C_1. Thus energy is supplied to the oscillator circuit and oscillations are started. The oscillating current in L_1 causes an alternating voltage of the same frequency to be induced in L_2. L_1 and L_2 are so connected that an increasing current through the valve and L_1 causes a voltage to be induced in L_2 which makes the grid positive. This voltage further increases the current through the valve until its maximum value is reached. When the current in L_1C_1 reverses, the voltage induced in L_2 causes the grid potential to fall. This reduces the current through the valve and further assists the oscillation in L_1C_1. Due to the amplification of the valve and the regeneration provided by L_1 and L_2, the amplitude of oscillation increases until steady operating conditions are reached. This occurs, as explained previously, when circuit losses, grid bias and the restricted positive grid voltage swing prevent oscillation from building up further.

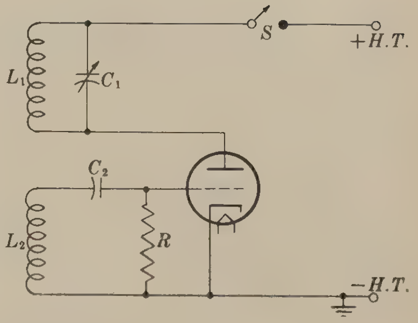


FIG.5-3 The series-feed tuned-anode oscillator.

When steady conditions have been reached, a circulating current flows in L_1C_1 at a frequency determined by the formula

$$f = \frac{1}{2\pi\sqrt{L_1C_1}}$$

Since the amplitudes of voltage and current are large in the anode circuit, high power can be drawn from the oscillatory circuit. Consequently the tuned-anode oscillator provides greater output power than the tuned-grid oscillator.

The circuit of Fig. 5-3 is called a *series-*

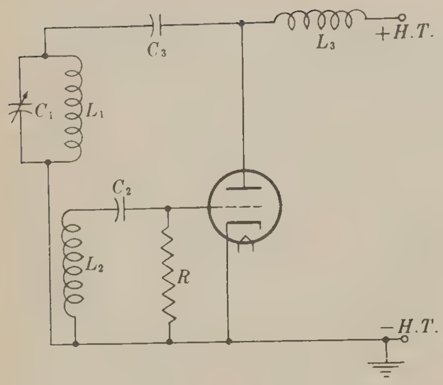


Fig.5-4 The parallel-feed tuned-anode oscillator.

feed tuned-anode oscillator, because the oscillatory circuit and the valve are connected in series with the H.T. supply. In this case the direct-current component through the valve also flows through the oscillatory circuit. Since the oscillatory circuit is at H.T. potential this arrangement is dangerous when tuning or coupling. Another disadvantage is that the a-c component of anode current flows through the H.T. supply. This results in increased a-c power losses in the H.T. supply and undesirable coupling to other circuits.

The circuit shown in Fig. 5-4 is designed to overcome these disadvantages. It is called a *parallel-feed* tuned-anode oscillator. The oscillator circuit and the valve are connected in parallel with the H.T. supply. The d-c component of anode current does not flow through the oscillatory circuit but is blocked from the oscillatory circuit by C_3. The alternating component of current through the valve flows through the oscillatory circuit L_1C_1 and is prevented from flowing through the H.T. supply by the choke coil L_3. Thus C_3 prevents the inductance L_1 from short-circuiting the H.T. supply while L_3 prevents the supply from short-circuiting the a-c voltage. The operation of this circuit is essentially the same as that of the series-feed oscillator. C_3 discharges somewhat when the current through the valve is a maximum. This supplies energy to the oscillatory circuit. During the remainder of the cycle the H.T. power supply charges C_3 through L_3. The values of inductance and capacitance in the oscillatory circuit are the chief factors in setting the frequency of oscillation. However, a vector analysis of the circuit shows that the frequency of oscillation is slightly different from the resonant frequency of the tank circuit when there is feedback and the circuit is oscillating. The following analysis assumes that the grid and anode currents and voltages are of sinusoidal waveform.

The oscillator is equivalent to a source of alternating e.m.f. connected in series with the anode resistance of the valve and the oscillatory circuit as shown in Fig. 5-5. This e.m.f. is μ times the grid excitation voltage E_g where μ is the amplification factor of the valve. Since these voltages are 180° out of phase with each other, the source of e.m.f. is represented by $-\mu E_g$. The negative sign indicates that the grid and anode voltages are 180° out of phase. The vectors representing these voltages form the reference vectors of the diagram in Fig. 5-6.

Since the grid excitation voltage is obtained by the inductive coupling between L_1 and L_2, the current I_L through L_1 must lag $-\mu E_g$ by 90°. The current

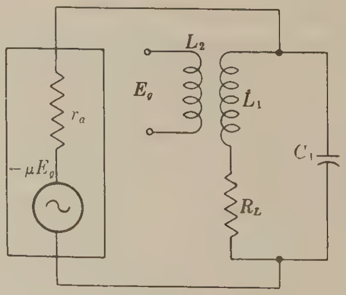


Fig.5-5 The equivalent circuit of an oscillator.

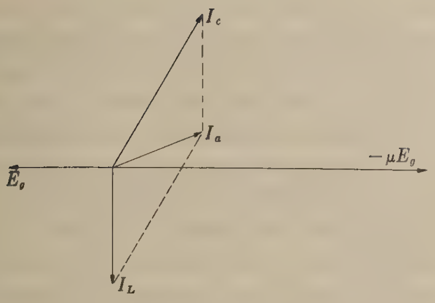


FIG.5-6 Oscillator vector diagram.

I_c through the capacitor is somewhat less than 180° ahead of I_L because of the resistance R_L of L_1. The vector sum of I_L and I_C is I_a, the a-c component of anode current. This anode current is observed to lead the source voltage $-\mu E_g$. That is, the oscillator circuit presents a capacitive reactance to the valve when there is positive feedback. The circuit is said to be *tuned capacitively* and oscillates at a frequency somewhat above the resonant frequency of L_1C_1.

5.3 Modifications for increased efficiency. The pulses of current which flow from cathode to anode through the valve may be analysed into two components; a d-c component, since current flows in only one direction through the valve and an a-c component of a complex nature which depends upon the operating conditions of the circuit. The fundamental frequency of the a-c component is the same as the frequency of oscillation of the circuit. The harmonics generated may vary widely in amplitude and are affected by such factors as the bias, the Q of the oscillatory circuit and the H.T. voltage. The circuit is adjusted to convert d-c power to a-c power as efficiently as possible. This requires that the d-c component of current through the valve be as small as possible relative to the a-c component. Two factors which are of prime importance in affecting the efficiency of an oscillator are impedance matching of the oscillatory circuit and the angle of current flow.

A. *Impedance Matching*

The output power which may be drawn from any source such as a battery, a generator, or a valve circuit is equal to the product of the current flowing from the source and the terminal voltage of the source. In order to draw a large current from the source, the impedance of the load must not be too great. On the other hand, if the impedance is reduced beyond a certain point a large proportion of the e.m.f. of the source is developed across the internal resistance so that the terminal voltage is reduced. The maximum product of voltage and current, that is, maximum output power is obtained when the impedance of the load equals the impedance of the source. When this condition is fulfilled, the two impedances are said to be *matched.*

In an oscillator, the valve is the source of a-c power and the LC circuit is the load. The impedance of a high-Q circuit at resonance is almost infinite as the voltage becomes very large and the supply current very small. In order to present a lower impedance to the valve, an anode tapping point is selected at some point X along the coil (Fig. 5-7). This tap is at a lower a-c voltage point on the coil than the end of the coil. Proper adjustment

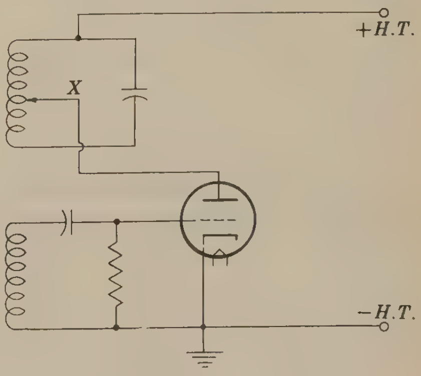


FIG.5-7 Adjustable anode tap for matching.

of this tap matches the impedance of the load circuit to the anode resistance of the valve and maximum power transfer to the oscillatory circuit occurs. This match is difficult to achieve in practice because the impedance of the valve source varies greatly throughout the cycle. The impedance at any instant equals the value of anode voltage divided by the value of current flowing through the valve. During the time the valve is cut off, the anode current is zero and so the impedance is infinite. As the grid voltage rises to a maximum, the anode voltage falls to a minimum. At the peak of the cycle the impedance becomes a minimum. The position of the anode tapping point which gives maximum power output is therefore found experimentally.

B. Angle of Current Flow

The fraction of a cycle during which current flows from cathode to anode in the valve is expressed in degrees and is called the *angle of current flow.* Its value is determined by the length of time in each cycle during which the grid is driven above the cut-off voltage. Fig. 5-8 shows the anode and grid voltages for an oscillator operating Class C. Since grid and anode voltages are 180° out of phase, the grid becomes most positive at a time when the anode is least positive. The anode voltage e_b is shown as a sinusoidal variation about the steady anode voltage E_b. The valve is biased beyond cut-off and the grid voltage e_c is a sinusoidal variation about this bias level. Anode current flows through the valve during that part of the cycle when the grid potential is above cut-off. Grid current flows when the grid is positive with respect to the cathode. The grid current flow maintains an automatic bias.

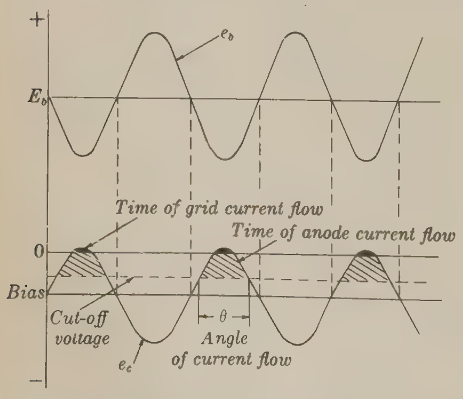


Fig. 5-8 Oscillator voltage waveforms, Class C operation.

The instantaneous power dissipated by the valve is equal to the product of the instantaneous anode voltage and the instantaneous anode current. The power dissipated during one cycle is the average value of the instantaneous power taken over one cycle. To make the efficiency a maximum, this power consumption must be made a minimum. This requires that the angle of current flow be kept to a minimum causing the current to flow as a large pulse of short duration when the anode voltage is a minimum. The grid therefore must be biased far beyond cut-off so that only the sharp peak of its positive voltage swing is effective in causing anode current to flow. Oscillators may be made to operate with 75% to 85% efficiency by reducing the angle of current flow to about 60 degrees.

The value of grid bias is determined for the most part by the size of the tickler coil, the amount of its coupling to the oscillating circuit coils and by the values of the grid-leak resistor and capacitor. In general, the larger the grid resistor and capacitor, the greater the bias developed and the more efficiently the circuit operates. However, an excessively large *RC* combination must be avoided as this may prevent the circuit from altering bias to follow changes in operating conditions. If the *RC* combination is too large the valve may be biased off completely for a number of cycles until the bias again becomes adjusted. This causes periods of oscillation each of which is followed by a period of no oscillation. This effect is termed *intermittent operation, motor-boating* or *squegging*.

C. Q of the Oscillatory Circuit

The Q of the LC circuit in an oscillator is usually set at about 12. A higher value of Q requires heavier construction of the inductor and capacitor to carry the large circulating current. If the Q is less than 12, energy losses are too large for efficient operation; the waveform of the output is no longer sinusoidal and the oscillator becomes unstable.

The frequency of an LC oscillator is usually adjusted by changing the value of C. To increase the frequency of oscillation, the capacitance must be reduced. If the frequency range is extended beyond certain limits in such a circuit, the change in Q of the circuit results in unstable oscillation. In order to make the oscillator cover a wide range of frequencies, the range is divided into frequency bands. A different inductor is used for each band and the variable capacitor provides continuous tuning within the band. This arrangement maintains a suitable value of Q over the complete range of frequencies.

D. Modified Tuned-Anode Oscillator

Partial tuning of the grid coil by means of a capacitor may be used to increase the efficiency of the tuned anode oscillator. A larger amplitude of alternating voltage is applied to the grid and consequently the bias is increased. A negative grid bias equal to twice the cut-off voltage may readily be obtained and the angle of current flow reduced to 120 degrees.

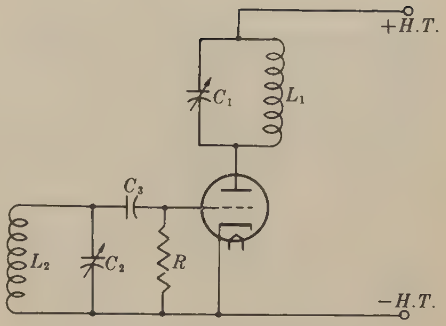


FIG.5-9 The tuned-anode tuned-grid oscillator.

5.4 The tuned-anode tuned-grid oscillator. The oscillator shown in Fig. 5-9 contains tuned LC circuits in both the anode and grid circuits. The coils L_1 and L_2 are so placed that there is no inductive coupling between them. Feedback from the anode to the grid circuit is provided by the interelectrode capacitance between the anode and grid within the valve. This capacitance is made to charge and discharge at the oscillation frequency by the changes in anode potential and the current flowing through the valve. This capacitance in turn supplies pulses of energy in the proper phase to maintain oscillation in the grid circuit. The anode and grid tuned circuits must offer inductive reactance to the flow of the alternating current in order to maintain the oscillation. Consequently the circuit oscillates at a frequency which is slightly lower than the resonant frequency of either circuit.

5.5 The Hartley oscillator. The oscillatory circuit L_1C_1 is connected between anode and grid in the Hartley oscillator shown in Fig. 5-10. The voltage across L_2, which is part of the oscillatory circuit, is applied between grid and cathode of the valve by means of C_2 and C_3. When the anode supply voltage is first applied the current increases from zero through the valve. This increasing current flows through the upper portion of L_1 and supplies energy to the oscillatory circuit. Oscillations are set up in L_1C_1 which develop an alternating voltage across L_1. This voltage is made up of two parts, the a-c component of anode voltage across

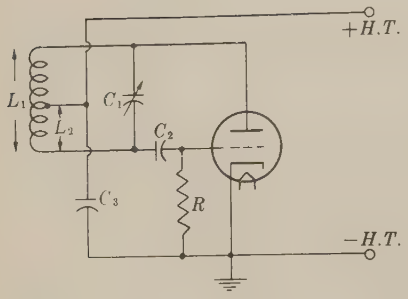


FIG. 5-10 The Hartley oscillator.

the upper section of L_1 and the a-c voltage across the lower part which is applied between grid and cathode of the valve to control the flow of current from the source. Since opposite ends of L_1 are in anti-phase the grid is made most positive when the anode is made least positive. This ensures that the feedback is positive and maintains oscillation. The tapping point on the coil is adjusted to provide the required grid excitation to the valve and to match the impedance of the oscillatory circuit to the impedance of the valve. Automatic bias is provided by C_2R to make the oscillator self starting. The circuit of Fig. 5-10 employs series feed as the oscillator circuit is at the same d-c potential as the supply battery. The capacitor C_3 completes the a-c circuit through the valve without short-circuiting the supply voltage.

A parallel-feed Hartley oscillator is shown in Fig. 5-11. Its operation is similar to the series-feed Hartley. The d-c blocking capacitor C_3 is placed on the anode side instead of the cathode side of L_1. The choke coil L_3 prevents the supply battery from short-circuiting the a-c output of the circuit and keeps a-c current out of the H.T. supply. The oscillatory circuit is at zero potential for d-c so that direct contact with the tuning capacitor is not dangerous.

Oscillators are sometimes operated with the anode at a-c ground potential, for example, in frequency-converter circuits in radio receivers. Such a circuit is shown in Fig. 5-12 and is called a *modified Hartley* circuit. The d-c path from ground is through the upper section of L_1, through the valve from cathode to anode, and through L_3 to the positive side of the H.T. supply. Hence, this circuit is series-feed. The anode is maintained at a-c ground potential by the by-pass capacitor C_4 and the d-c blocking capacitor C_3. The grid and cathode are therefore at high a-c potential instead of the grid and anode. Otherwise the operation of this circuit is the same as described for the conventional Hartley oscillator. It should be noted that any one of the three electrodes, cathode, grid or anode, may be kept at a-c ground potential provided the other two are left free to assume a-c potentials independent of each other and of ground.

The Hartley circuit oscillates at a slightly lower frequency than the resonant frequency of the tuned circuit. This means that the load is inductive for positive feedback.

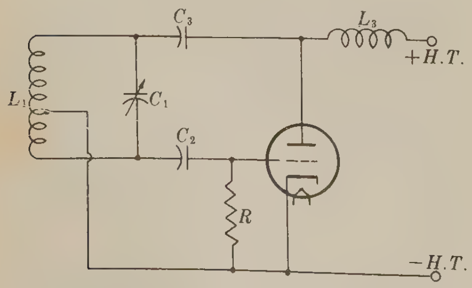


FIG. 5-11 The parallel-feed Hartley oscillator.

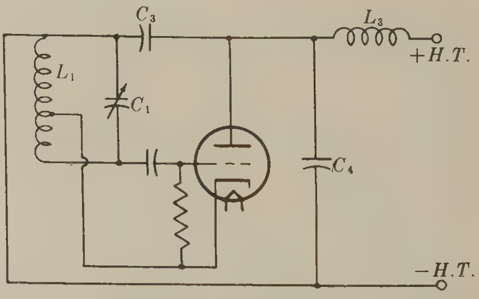


FIG. 5-12 The modified Hartley circuit.

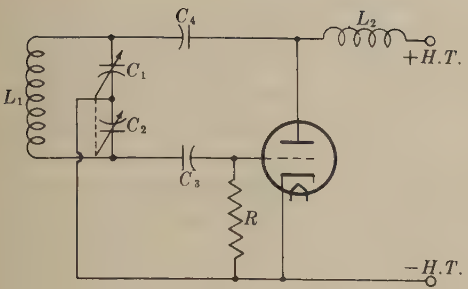


FIG.5-13 The Colpitts oscillator.

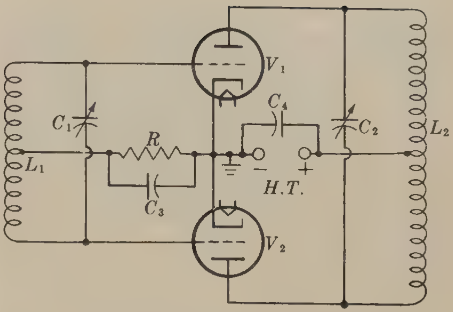


FIG.5-14 The push-pull oscillator.

5.6 The Colpitts oscillator. The Colpitts oscillator (Fig. 5-13) is similar to the parallel-feed Hartley circuit with the exception that the feedback voltage is obtained from a section of the tuning capacitor instead of the inductor. The tuning capacitor is divided into two sections C_1 and C_2 with the rotors mechanically coupled or *ganged.* The alternating voltage developed across C_2 is applied between the grid and cathode through C_3 to provide the feedback. This circuit oscillates at a frequency slightly higher than the resonant frequency of the tuned circuit. This means that the load is capacitive for positive feedback.

5.7 Push-pull oscillator. A push-pull oscillator is similar in design and construction to a push-pull r-f amplifier. The tuned-anode tuned-grid oscillator shown in Fig. 5-14 consists essentially of two triode valves with one oscillatory circuit connected between the grids and one between the anodes. The capacitor C_3 maintains the centre point of the tuned grid circuit at a-c ground potential so that the voltages applied to the two grids are always 180° out of phase. The tuned anode circuit is similarly balanced to ground by the by-pass capacitor C_4.

When the filaments are heated and the H.T. supply is completed, current flows from ground through each valve from cathode to anode and through L_2 to the positive side of the H.T. supply. When the circuit is first turned on, it is highly improbable that the two valves will continue to conduct the same amounts of current at every instant. Any slight difference in conductivity of V_1 and V_2 will cause oscillation to begin in the tuned anode circuit. The potential of one anode therefore tends to rise above the supply potential while that of the other anode falls below the supply potential. This causes unequal charging of the interelectrode capacitances between grid and cathode which in turn sets up oscillations in the grid circuit. Sufficient energy must be fed back from the anode to the grid circuit to compensate for losses in the grid circuit in order to maintain oscillation. Triode valves are usually used as their large interelectrode capacitance provides the necessary feedback of energy. Oscillation developed in the tuned grid circuit causes one grid to become positive while the other grid is made negative. One valve therefore conducts heavily while the other is driven to cut-off. That is, two pulses of energy are supplied to each oscillatory circuit per cycle. The d-c source is thus made to supply energy to the tuned anode circuit at the frequency of the grid-circuit oscillations. The anode circuit is tuned slightly off resonance to present a capacitive load to the valves for positive feedback.

Automatic grid-leak bias is developed across the grid-leak resistor R. If the circuit is operated at sufficiently high frequency, the grid-cathode capacitance of the valves is large enough so that C_3 can be dispensed with. Sufficient bias is developed to make the oscillator operate Class C so that high efficiency is attained. The output is usually obtained by transformer coupling to the tuned anode circuit.

Several advantages are to be derived from the use of the push-pull circuit for oscillators. The power that can be produced is approximately twice that for a single valve. Since the interelectrode capacitances are effectively in series, they are less likely to cause *parasitic oscillations* than two valves in parallel. Parasitic oscillations are extra oscillations at frequencies different from those desired. The presence of parasitic oscillations results in reduction of output power and erratic operation. The symmetry of the circuit permits the use of short lead wires and rigid mountings, a decided advantage in high-frequency circuits. Finally, the even harmonics are not produced by a push-pull oscillator. This makes it possible to operate the circuit at high efficiency and maintain a good output waveform.

5.8 Frequency stability. It is of the utmost importance that the frequency of the r-f waves from a radio transmitter remain constant. Any variation from the assigned frequency makes it difficult for the radio receiver to be kept tuned to the transmitter frequency and also interferes with the reception from transmitters on neighbouring frequencies.

The causes of *frequency drift* or *instability* in a transmitter are

(i) physical changes in the components of the oscillator of the transmitter

(ii) electrical changes in the oscillator.

Physical changes include the variations in size or shape of circuit components due to faulty mounting, vibration and heating effects. Any change in the dimensions of the inductor or capacitor in the tuned circuit results in a shift of frequency of oscillation. For this reason, the inductors must be rigidly mounted to prevent any change in the spacing between turns. The inductor is made of conducting wire or tubing large enough to carry the current without developing appreciable amounts of heat. Air capacitors in the oscillator should be firmly mounted so that the spacing between the plates remains constant during normal operation. If heating of capacitors which have other than air dielectric cannot be avoided, capacitors with low temperature coefficients should be used. It is common practice in high precision equipment to enclose the entire oscillator valve and its associated circuits in a temperature-controlled oven to maintain constant frequency.

Causes of frequency drift due to electrical changes include changes in the supply voltages to anode and filaments, changes in valve characteristics and in loading. The first of these causes may be eliminated by using a power supply equipped with voltage regulators which maintain a constant output voltage when the input mains voltage varies. Small receiving valves are used in preference to the larger transmitting types in oscillators because of their smaller elements and more rigid mounting of electrodes. More serious is the variation in frequency of oscillation caused by changes in the load impedance. For this reason an oscillator is very seldom coupled directly to an aerial. A valve amplifier called a *buffer* is connected between the oscillator and the aerial. The buffer presents a constant impedance to the oscillator as well as amplifying its output. Thus variations in aerial impedance do not change the impedance presented to the oscillator. Such a transmitter is referred to by the name *master-oscillator power-amplifier* transmitter.

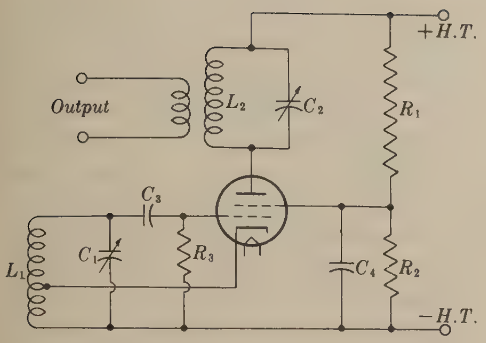


FIG.5-15 An electron-coupled oscillator.

5.9 The electron-coupled oscillator. Frequency drift caused by loading the oscillator can be avoided to a large extent by using an *electron-coupled oscillator* shown in Fig. 5-15. The valve is a tetrode in which the cathode, control grid and screen grid are used in a modified Hartley oscillator. The screen of the valve, operating as the anode of the oscillator, is maintained at a-c ground potential by the by-pass capacitor C_4. R_1 and R_2 form a voltage divider which supplies a constant d-c voltage to the oscillator anode. The frequency of oscillation is determined by the values of L_1 and C_1.

The tank circuit L_2C_2 is the load circuit for the oscillator. It is coupled to the oscillator only by the electron stream passing through the valve and has little effect on the frequency of oscillations. L_2C_2 receives pulses of current at the oscillator frequency. When tuned to resonance at this frequency, a large circulating current is set up in L_2C_2. The output power is taken from this tank circuit.

Pentode valves are also used in electron-coupled oscillators. The suppressor grid is connected to ground and acts as additional shielding between the oscillator section and the output section of the valve. Thus the effect of the output circuit on the frequency of oscillation is further reduced.

5.10 The piezo-electric effect. The frequency stability of an oscillator can be made extremely high by replacing the *LC* tuned circuit by a crystal. The use of crystals for this purpose depends upon the *piezo-electric effect.*

This effect is exhibited by certain crystalline substances, principally quartz, Rochelle salt and tourmaline. Of these, quartz is most commonly used because of its availability and durability, though it is not as sensitive as the other twc types. A crystal of quartz is shown in Fig. 5-16 and a geometrical representation in Fig. 5-17. The diagonals of the crystal, Fig. 5-17 (a), are called the *electrical axes* X_1, X_2, *and* X_3. Between these are the *mechanical axes* Y_1, Y_2 *and* Y_3. The vertical axis of the crystal is called the *Z*-axis. A crystal section cut so that its flat sides are perpendicular to an electrical axis is called an *X-cut* crystal, Fig. 5-17 (b). A section whose sides are cut perpendicular to a mechanical axis is the *Y-cut* crystal.

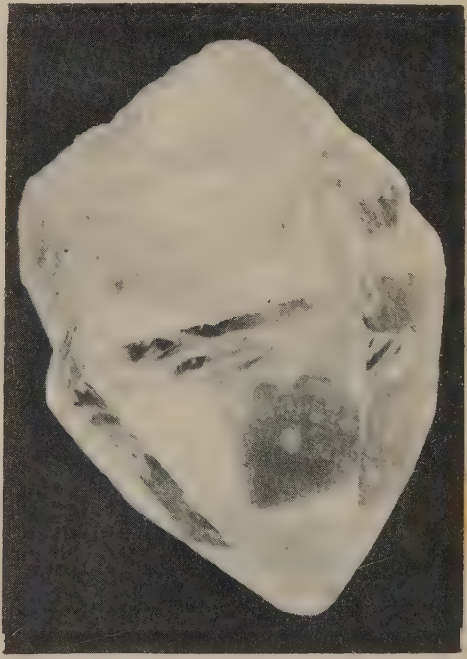
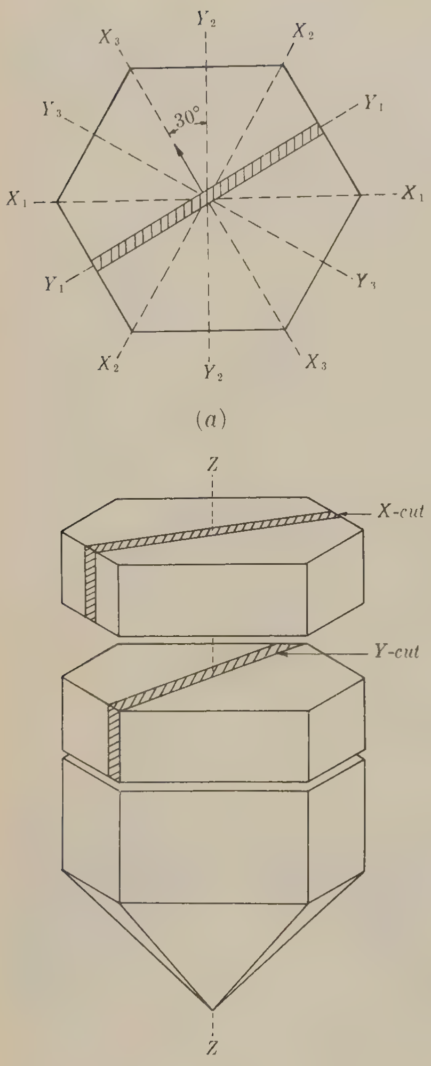


FIG.5-16 A quartz crystal.



(a)

(b)

Fig. 5-17 Geometrical representation of a quartz crystal showing (a) cross-section and axes (b) full crystal and X and Y cuts.

If a mechanical pressure is exerted across the crystal in the direction of its mechanical axis, a p.d. is developed in the direction of the electrical axis. If the mechanical force is one of tension instead of compression, the polarity of the p.d. is reversed. Conversely if a p.d. is applied to opposite ends of the electrical axis, a mechanical strain is produced along the mechanical axis. The polarity of the p.d. determines whether the strain is an expansion or a contraction.

If an alternating voltage is applied across the faces of the crystal, the mechanical strains set up cause the crystal to vibrate. The amplitude of these vibrations is very small at most frequencies but becomes quite large at one frequency. This frequency is the *mechanical resonant frequency* of the section. It depends upon the dimensions of the section, particularly its thickness. The temperature of the crystal affects the resonant frequency to a lesser degree. At this frequency, the amplitude of oscillations may become so large as to shatter the crystal.

A crystal section responds to an applied a-c voltage in the same way as a series *LCR* circuit and may be regarded as such electrically. The equivalent inductance, capacitance and resistance values can be measured or calculated. Typical values for a section of quartz 0.636 cm. thick in the direction of vibration are as follows:

Resonant frequency = 451 Kc/s
Equivalent inductance = 3.65 henries
Equivalent capacitance = .0316 micro-microfarads
Equivalent resistance = 9040 ohms.

The value of Q for the equivalent series circuit is extremely high, about 1100 for this example. It is this property of a crystal which makes it useful for maintaining a constant frequency in an oscillator.

It is to be emphasized that the resonant frequency of a crystal section depends almost entirely upon its dimensions, particularly its thickness in the direction of vibration. The frequency is inversely proportional to this dimension. Crystal sections vary in thickness from 5.7 mm. at about 500 Kc/s to 0.14 mm. at 20 Mc/s. At this upper limit the crystal becomes extremely fragile and care must

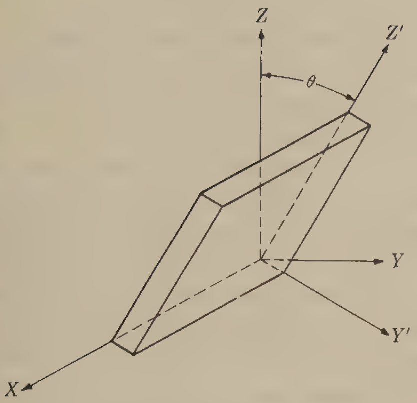


Fig.5-18 An AT-cut crystal.

be taken to prevent the voltage across it from shattering the crystal.

Temperature affects the resonant frequency to a minor extent. To eliminate frequency variation due to temperature change, the crystal sometimes is enclosed in an oven the temperature of which is thermostatically maintained at a constant value. It is possible however to cut the section from the whole crystal in such a way that the change in frequency due to change in temperature is almost zero. Such a crystal section shown in Fig. 5-18 is called an *AT-cut* crystal and provides a constant frequency even without a temperature-controlled oven.

In addition to the *X*, *Y* and *AT* cuts, many other cuts are shown in the photograph of Fig. 5-19 (a). These cuts differ in the angles which they make with the electrical and mechanical axes of the crystal. As a result they have different thickness-frequency relationships and different frequency-temperature coefficients. For example, *X*-cut crystals have coefficients of from −15 to −25 cycles per megacycle per centigrade degree. The negative sign indicates that as the temperature increases, the resonant frequency decreases. For *Y*-cuts the coefficient ranges from −20 to −100 cycles per megacycle per centigrade degree.

Of the crystal materials which may be used, quartz is the most widely employed because of its durability. Rochelle salt exhibits the piezo-electric effect to a much greater degree than quartz but is very fragile. Tourmaline is a rare substance and therefore expensive. Even with quartz, certain precautions are necessary. The faces should be very clean. For this reason and for protection from damage the crystal is enclosed in a holder. Fig. 5-20 and Fig. 5-21 illustrate two types of holder in common use.

In the holder of Fig. 5-20 the crystal section is held firmly between metal plates or electrodes and enclosed in a bakelite case. External connections to the two plates are led through the bottom

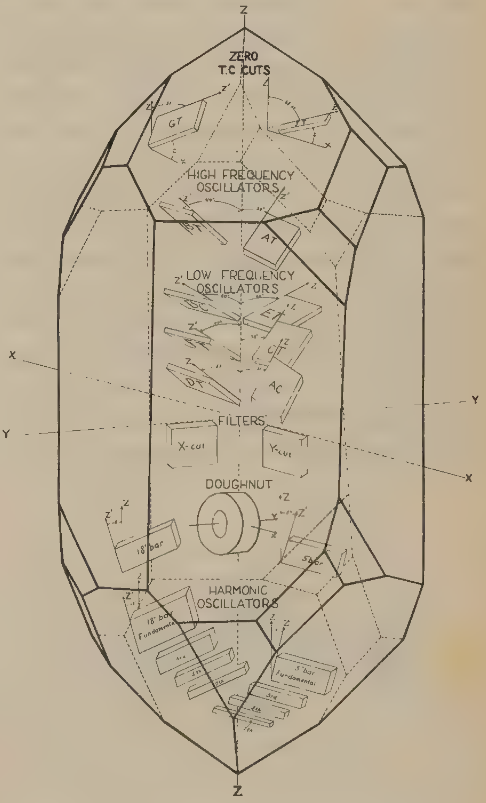


Fig.5-19 Different crystal cuts and their uses.

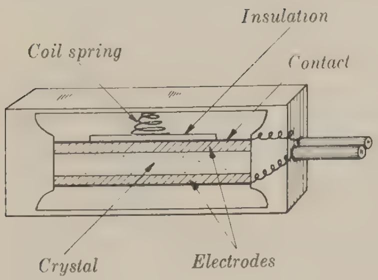


FIG. 5-20 Fixed-frequency holder.

of the case. Crystals are often mounted so that a small air gap exists between one metal plate and the face of the crystal as in Fig. 5-21. The resonant frequency depends somewhat on the thickness of the air gap. Adjustment of this gap makes it possible to alter the resonant frequency by as much as 1 Kc/s per Mc/s. *X*- and *Y*-cut crystals are not suitable for this type of operation as the frequency does not vary smoothly with thickness of the gap. The frequency tends to jump abruptly as the air gap is adjusted. The capacitance between the crystal holders is effectively in parallel with the equivalent circuit of the crystal. The capacitance of the air gap is electrically in series with the equivalent circuit. These are shown in the diagram of Fig. 5-22 (a). The circuit symbol for a crystal is shown in Fig. 5-22 (b).

Crystals are used in two types of oscillator circuits. These are known as the *Miller* and the *Pierce* crystal oscillators.

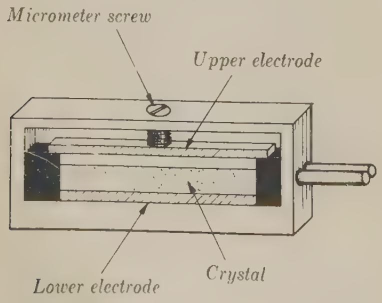


FIG. 5-21 Adjustable-frequency holder.

5.11 The Miller crystal oscillator. The diagrams of a simple Miller crystal r-f oscillator and its equivalent circuit are shown in Fig. 5-23 (a) and (b). The crystal is connected between grid and cathode of the valve as in a tuned-grid oscillator. The anode load is a tuned *LC* circuit. Feedback from the anode to the grid circuit is provided by the interelectrode capacitance between these two electrodes. At low frequencies where this capacitance is not sufficient to maintain oscillation, an external capacitance may be connected from the anode to the grid. Automatic bias is provided by the grid leak *R* and the equivalent capacitance of the crystal.

Since the Miller circuit is a tuned-grid oscillator, the anode circuit must present an inductive load to provide positive feedback. Therefore the oscillatory circuit must be tuned to a higher frequency than the crystal frequency. The effect which the tuning of L_1C_1 has on the amplitude of oscillations is shown in the graph of Fig. 5-24. If C_1 is made very small, the circuit is tuned to a high frequency and presents an inductive load. As C_1 is increased, the resonant frequency of L_1C_1 approaches the frequency of the crystal and the ampli-

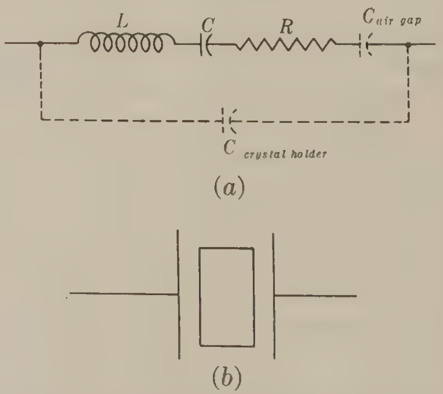


FIG. 5-22 The circuit symbol and equivalent circuit of a crystal.

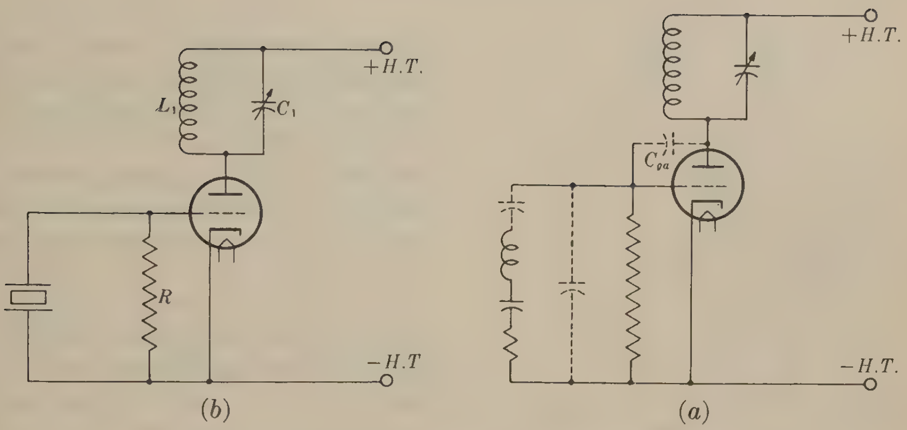


FIG.5-23 The Miller crystal oscillator and its equivalent circuit.

tude of oscillation increases. If the circuit is tuned below the crystal frequency, the load becomes capacitive. The feedback is then degenerative and oscillation ceases abruptly. The dotted line indicates the correct tuning of the oscillatory circuit. It should be set close to the crystal frequency but far enough from it to ensure that slight changes of loading, temperature and operating voltages do not cause oscillation to stop.

The amplitude of oscillation is indicated most readily by a milliammeter connected in the cathode circuit of the valve. As the amplitude of oscillation increases, the automatic grid-leak bias increases causing the steady anode current component to decrease. At maximum oscillation the d-c anode current becomes a minimum (Fig. 5-25). The value of C_1 should be adjusted to give minimum d-c current through the valve and then decreased slightly to ensure stability of operation.

An r-f choke coil may also serve in place of the oscillatory circuit in a Miller oscillator. It offers an inductive load to the valve if properly designed and with its self-capacitance forms the anode tuned circuit.

The output power of a crystal oscillator is usually small due to the fragility of the crystal. Coupling is made from the anode of the valve and should be loose to avoid overloading. If greater power is required, the oscillator may be followed by an amplifier or a modified Miller circuit may be employed. Fig.

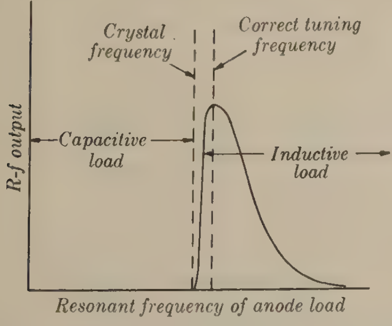


FIG.5-24 Frequency response over tuning range of Miller oscillator.

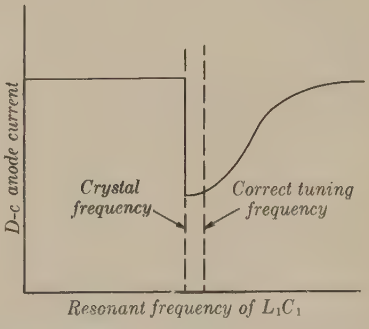


FIG.5-25 Anode current over tuning range of Miller oscillator.

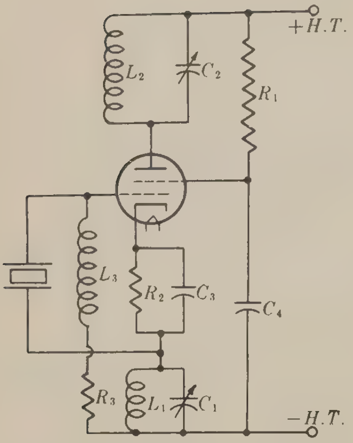


FIG.5-26 Modified Miller circuit.

5-26 shows a modification using a beam power valve that provides up to 15 watts of output from an electron-coupled circuit. This circuit is sometimes called a *tri-tet* oscillator.

The screen grid serves as anode of the oscillator and is kept at r-f ground potential by the by-pass capacitor C_4. The tuned circuit L_1C_1 is therefore placed in the cathode circuit and is tuned above the crystal frequency to present an inductive load. Bias is provided by R_3 and the capacitance of the crystal holder. In addition the r-f choke L_3 is included in the grid circuit to increase the r-f impedance between grid and ground. This prevents the relatively low impedance of R_3 from shorting the r-f voltage at the grid. In addition to the automatic bias, some cathode bias is provided by R_2C_3. This prevents the valve from being damaged by excessive current in the event that oscillation ceases and automatic bias fails. The output circuit L_2C_2 is electron coupled to the crystal oscillator section. This electron coupling isolates the crystal circuit from the final load so that the loading in the anode circuit has not so much effect on the frequency as in the simple Miller circuit.

5.12 The Pierce crystal oscillator. In the Pierce oscillator, Fig. 5-27 (a), the crystal is connected between the grid and anode. C_1 is a blocking capacitor in series with the crystal to protect it from the high d-c voltage. Bias is provided in the cathode circuit by R_2C_3 and in the grid circuit by R_1 and the capacitance of the crystal holder. The Pierce circuit is the equivalent of the Colpitts oscillator as indicated by the equivalent circuit diagram of Fig. 5-27 (b). Feedback is provided by the r-f voltage developed across the grid-cathode capacitance. For operation at low frequencies it is some-

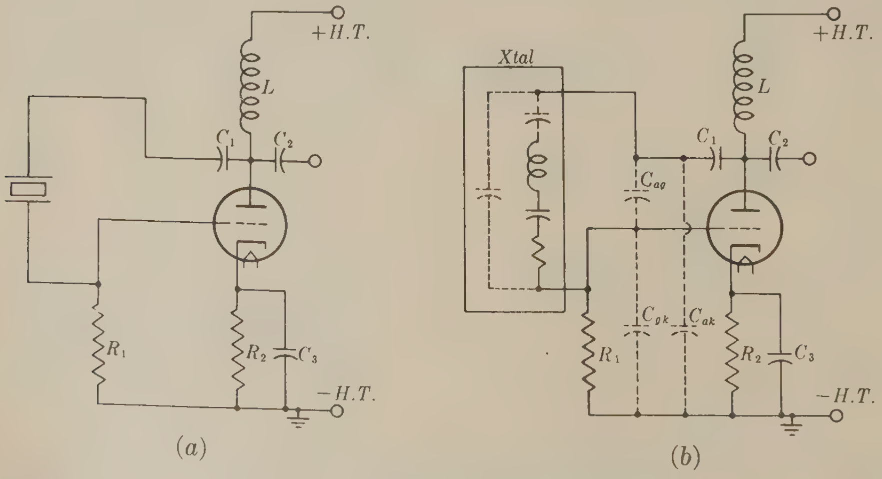


FIG.5-27 The Pierce crystal oscillator and equivalent circuit.

times necessary to add an external capacitor in parallel with C_{gk}. This overcomes sluggishness and makes the oscillator quick-starting for keying purposes.

The load inductor L, together with its self-capacitance and interelectrode capacitance, must present a capacitive reactance to the valve to provide positive feedback.

The power output of the Pierce oscillator is limited to a few watts at most. The r-f current flowing through the crystal must be small to prevent heat developing and causing the frequency to drift. The voltage across the crystal must also be kept moderately low to avoid fracture. These factors limit the use of the crystal oscillators to heterodyne oscillators in receivers, frequency standards in test equipment and master oscillators in transmitters. When high r-f power is required at constant frequency, an amplifier is used following the crystal circuit.

Crystal oscillators are much less flexible than tuned-circuit oscillators. A separate crystal must be connected in the circuit for each desired frequency. Below about 500 Kc/s the crystals become heavy and sluggish and their use becomes impractical. The upper limit of oscillations is about 20 Mc/s, at which the thickness is in the neighbourhood of 0.14 mm. This limit may be extended however by connecting one or two stages of frequency doublers following the crystal oscillator. In this way good frequency stability may be maintained to about 80 Mc/s.

5.13 Very-high-frequency oscillators. As the frequency of an oscillator is raised, the radio-frequency losses increase and the output power available becomes less. Some of these losses are due to the external circuit of the oscillator and some to the limitations of the valve. The following section is devoted to a discussion of the reduction in efficiency of oscillator circuits at very high frequencies and the modifications required to overcome the defects.

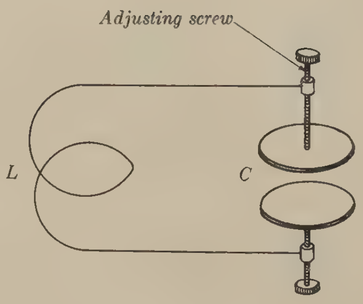


FIG.5-28 High-frequency LC circuit.

A. Losses Associated with Oscillatory Circuits

In order to increase the frequency of oscillation it is necessary to decrease both the inductance and capacitance of the oscillatory circuit. This procedure is extended until the inductor is reduced to one or two turns of heavy wire and the capacitor consists of two parallel metal plates several millimeters apart. Fig. 5-28 shows an LC circuit used to generate oscillations in the neighbourhood of 200 Mc/s. At these frequencies the Q of the circuit falls off badly and so the efficiency of the oscillator decreases. Radiation from the circuit becomes very large due to the rapid reversals of current and the open construction of the coil and capacitor. This has the effect of introducing resistance into the tuned circuit which reduces its effective Q.

Skin effect also becomes a determining factor. The high-frequency current is forced to flow in a thin layer on the surface of the conductor. The thickness of the conducting layer is inversely proportional to the square root of the frequency. For example, at 200 Mc/s the conducting layer is only $\frac{1}{10}$ as thick as it is at 2 Mc/s. Skin effect therefore, sets up a greater resistance to the flow of current which further decreases the Q

of the circuit. This defect can be reduced to some extent by using conductors of large diameter to expose a greater surface area. Plating the conductor with a metal of low resistivity such as silver also helps to reduce losses due to skin effect.

Eddy currents induced in neighbouring conductors by the large currents flowing in the *LC* circuit also drain energy from the circuit. These currents produce heat which represents a loss in power from the oscillator. The reduction in efficiency makes the use of *LC* circuits prohibitive for very-high-frequency generators.

B. *Transmission-Line Oscillators*

If an alternating current is made to travel along a wire conductor called a *transmission line*, two points a half wavelength apart will be 180° out of phase with each other. This is precisely the phase relationship between the two ends of an oscillatory circuit. A half-wave conductor bent in the form of two parallel sections each one quarter wavelength long, shorted at one end, is referred to as a *quarter-wave line*. Such a line contains inductance in each element of its length and capacitance between the two sections. It is electrically equivalent to a high-Q resonant circuit. Since at any instant the current travels in opposite directions in the two sections, the magnetic fields about the two conductors tend to cancel each other. By placing the sections close together the loss of energy due to radiation and eddy currents may be kept to a very low value. The sections are usually made of metal tubing of comparatively large diameter to reduce the effective resistance due to skin effects. Finally, the sections are connected directly to the grid and anode terminals of the valve in push-pull circuits reducing the length of leads to a minimum.

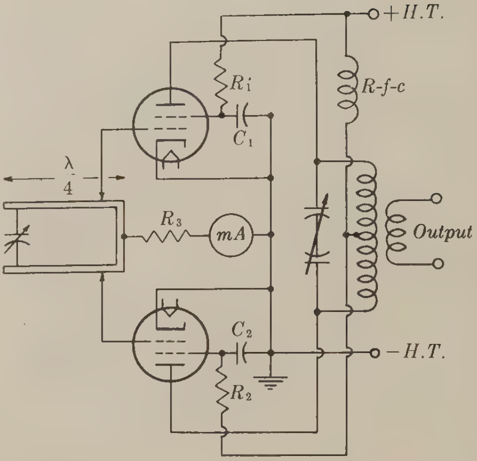


Fig. 5-29 Push-pull transmission-line oscillator.

Tuning of the transmission line is accomplished by adjustment of a shorting bar across the two parallel sections. If the bar is moved closer to the open end, the wavelength generated is reduced and hence the frequency of oscillation is increased. A small two-plate capacitor placed across the open end is also effective in tuning the circuit.

A transmission-line oscillator of the push-pull type is shown in Fig. 5-29. Heater circuits have been omitted as they are quite conventional. The circuit employs a dual beam power valve and is designed to operate at 120 Mc/s. The tuned grid circuit consists of two parallel sections of $\frac{3}{8}$-inch copper tubing $16\frac{1}{2}$ inches long spaced $1\frac{3}{4}$ inches between centres. The lines are shorted at one end by a fixed copper strap and tuned by a small trimmer capacitor at the open end. They are mounted $1\frac{1}{4}$ inches from the base board on ceramic stand-off insulators. The grids are tapped about $2\frac{3}{4}$ inches from the short-circuited end of the line. By tapping at a low voltage point the grid losses are kept small compared with the circulating energy in the transmission line. This maintains a high Q and provides good frequency stability of the oscillator.

The tuned anode circuit consists of a small coil and split-stator capacitor. The coil consists of two turns of No. 12 enamelled wire wound with a $\frac{3}{4}$-inch inner diameter and $\frac{1}{2}$-inch spacing. The supply voltage is applied through an r-f choke to the centre of the coil and hence to the two anodes. The screen voltages are obtained through 25,000Ω 10W voltage dropping resistors R_1 and R_2. The screens are by-passed to ground by .002μF capacitors C_1 and C_2. Automatic bias is developed by means of the 15,000-ohm, 1-watt resistor R_3 and input capacitance of the valves. Since the screen reduces the grid-to-cathode capacitance to .2 micromicrofarads, there is insufficient internal capacitance to maintain oscillation. Additional feedback is obtained by running a short length of wire from each anode close to the corresponding grid line to form a feedback capacitor. The capacitance between the feedback line and the grid circuit is adjusted by clipping and bending the line to give the correct grid excitation. Transformer coupling to the load circuit is provided by two turns of No. 12 enamelled wire closely wound with $\frac{3}{4}$-inch inner diameter. The coupling may be varied by adjusting the position of this coil relative to the anode tank coil.

The oscillator is adjusted for operation by the following steps. First, the grid-circuit trimmer capacitor is adjusted to provide the correct oscillator frequency. Following this, the feedback capacitors are adjusted with a $0-10mA$ d-c meter connected between the grid-leak resistor and ground. Proper grid excitation is obtained when the grid current is between 3.5 and $5mA$. Lastly the anode circuit is tuned for maximum efficiency. A 25-watt lamp connected across the output in place of the aerial may be used to indicate the correct setting of the anode tuning capacitor and coupling. With 400 volts supplied to the anode, the circuit may be loaded until the anode current reaches $150mA$ when properly tuned. An output of 60 watts is possible from such an oscillator. It is suitable for c-w, m-c-w or r-t transmission. Thirty watts output from a modulator connected into the anode circuit is sufficient to modulate the signal fully.

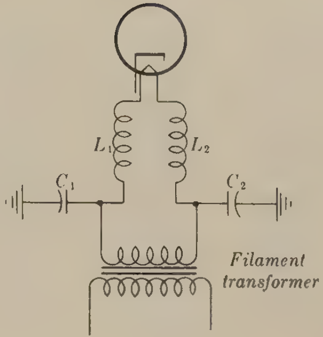


FIG. 5-30 Isolating the filament transformer from r-f current.

A number of variations of the circuit of Fig. 5-29 are possible. The transmission-line circuit may be used in the anode circuit or in both anode and grid circuits. Such an oscillator does not require the use of the push-pull circuit. Single-valve oscillators are in common use with tuned transmission lines in anode, grid and cathode circuits.

C. Cathode Circuits

When the circuit requires that the cathode be at high r-f potential, precautions must be taken to prevent the r-f current from flowing in the secondary of the filament transformer. The low reactance between cathode and heater makes it impossible to maintain one at an r-f potential and not have the other at the same r-f potential. Fig. 5-30 shows one method of preventing r-f current from entering the filament transformer. L_1 and L_2 are r-f choke coils which offer high impedance to r-f current. C_1 and C_2 are by-pass capacitors which maintain the secondary of the

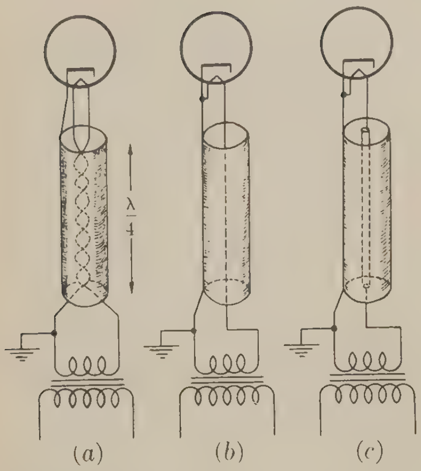


FIG. 5-31 Quarter-wave transmission-line cathode circuits.

filament transformer at r-f ground potential. Their capacitance is so small that there is very little shunting effect produced on the low-frequency heater current.

More frequently, quarter-wave transmission lines are used in the cathode circuit to keep the heater at high r-f potential. Three common methods of supplying heater current in oscillators are shown in Fig. 5-31. Each employs the principle that one quarter wavelength along a transmission line from a point of maximum impedance there is a point of minimum impedance. Consequently a quarter-wave transmission line may be grounded at one end and be at high r-f potential at the other end. In Fig. 5-31 (a) the heater leads are brought out through the tubular cathode line. The filament transformer is grounded and the heater and cathode, a quarter wavelength away, are at a high r-f potential. In (b) the cathode line is used as one side of the heater supply, the other side being a separate conductor through the cathode line. Concentric tubes are used to form the conductors to supply the heater current in (c). These are insulated from each other at the power supply frequency but may be considered as a single conductor at the transmitter frequency due to the low impedance between them.

Some circuits require that the filaments be at r-f ground potential in spite of the large reactance of the filament-lead inductance at high frequency. The circuit of Fig. 5-32 shows how this inductance may be included as part of a tuned line. The line is a half wavelength long from the filament to the grounded end. Since this half-wave line presents the same impedance to ground at both ends, the effect is the same as grounding the filament itself inside the valve.

D. *Concentric-Line Oscillators*

A quarter-wave section of concentric transmission line may be used to replace the open-line tuned circuits described in part B. A tuned-grid tuned-anode oscillator which contains a concentric-line grid circuit is shown in Fig. 5-33. Because of the very high Q of the concentric line the frequency stability is high, comparable to crystal control over 10 to 20 Mc/s. The inner and outer conductors are short-circuited at the cathode end. Therefore they show a high impedance between them at the upper end which is one quarter wavelength distant. The grid is tapped down on the line for loose coupling in order that grid losses be small compared to the power in the concentric line. This prevents excessive

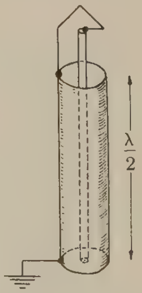


FIG. 5-32 A half-wave transmission-line cathode circuit.

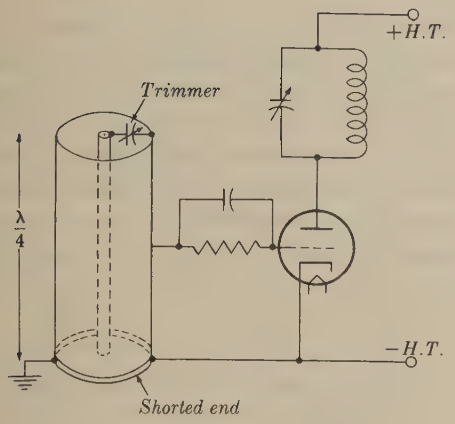


FIG.5-33 Tuned-grid tuned-anode transmission-line oscillator.

loading of the concentric line and maintains its high Q. The concentric-line circuit has one additional advantage over the open-wire type. The outer surface of the outer conductor is at ground potential and serves to shield the r-f current. The field is confined inside the line. This reduces losses due to radiation. Tuning the concentric line presents a greater difficulty however. Its effective length is changed by adjustment of a sliding plunger which shorts the inner and outer conductors. The contact must be maintained clean and firm to keep the resistance to current flow a minimum. Fine control of frequency may be effected by the use of the trimmer capacitor across the open end of the line.

5.14 Ultra-high-frequency oscillator valves. The preceding section is devoted to a description of the defects of conventional oscillator circuits at high frequencies and the modifications required to extend the frequency range. Special attention must also be given to the valves used to generate power at very high frequencies. The upper limit of oscillation frequency for ordinary small receiver valves is in the region of 150 Mc/s. Valves with reduced interelectrode capacitance are used in high-power transmitters up to 300 Mc/s. Oscillations as high as 1700 Mc/s may be generated by using valves of very special construction connected to external tank circuits but the useful power output becomes extremely low. For the generation of higher frequencies the oscillatory circuit is incorporated inside the glass envelope as a part of the valve itself. The following paragraphs deal with the limitations which valve construction imposes on the frequency of operation.

A. Inductance and Capacitance Associated with the Valve Elements

In r-f oscillators operating at high frequencies up to a megacycle per second the capacitance of the oscillatory circuit is so large compared with the interelectrode capacitance of the valve that the latter is usually neglected. As the frequency is increased the capacitance of the tank circuit must be decreased so that eventually the capacitance between the valve electrodes limits the frequency of the oscillator. Moreover, the lead wire to each electrode inside the glass envelope is equivalent to a small inductance in series with that electrode. Consequently there is a limit to which the frequency may be raised by decreasing the inductance of the tank circuit. In the extreme case, the external inductance is reduced to that of the grid and anode leads. Oscillations are generated at the resonant frequency of the valve. To increase the frequency further, the valve must be constructed to minimize the interelectrode capacitances and the lead inductances.

Low interelectrode capacitance may be attained by reducing the size of the electrodes and spacing them farther apart. This practice has its disadvantage, however. The reduction in size of the elements decreases the ability of the valve to dissipate heat and therefore limits the power output. Spacing the electrodes farther apart inside the valve makes it necessary to use higher voltages if the electron transit time is not to be increased.

The leads of high-frequency valves are brought out at widely separated points to reduce interelectrode capacitance in the valve. This construction is illustrated by the doorknob and acorn valves shown in Fig. 1-108 and Fig. 1-109. The bakelite base is omitted from these valves to eliminate dielectric losses as much as possible. To reduce lead inductance within the glass envelope, double connections are sometimes made to the anode and grid. Each electrode has two terminals which are symmetrically connected to the external circuit so that their lead inductances are in parallel making the effective inductance less.

B. Electron Transit Time

In low-frequency circuits, the length of time required for an electron to pass from cathode to anode is a very small fraction of the a-c period. The space current flowing through the valve is in phase with the grid voltage controlling that current at every point between the electrodes. As the frequency is increased, the duration of a cycle becomes less. Eventually the time of flight of an electron from cathode to anode becomes an appreciable fraction of a complete cycle. When the electron transit time becomes $\frac{1}{10}$ of a period or greater, the efficiency of the oscillator begins to fall off. The reason for this is as follows.

In a low-frequency r-f oscillator when the anode current reaches its peak value, the voltage is a minimum. The instantaneous power dissipated at the anode is equal to the product of the anode current and the low anode voltage. The average anode dissipation may be kept to within about 15 percent of the total power converted from d-c to r-f power so that the oscillator is maintained at high efficiency. If, due to electron transit time, the anode voltage lags the grid voltage, the maximum current through the valve will flow at a time before the anode voltage reaches its minimum value. Consequently the product of current and voltage is increased. The average power dissipated at the anode is therefore greater and the efficiency of the oscillator is reduced.

C. Induced Grid Current

When an electron, which is a negative charge, approaches a grid, it repels electrons along the grid wires. As the electron recedes from the grid this force disappears and electrons return to the grid wires. When an a-c is applied to the input of the valve there is a gradually increasing and decreasing current through the valve. Thus the numbers of electrons in the region of the grid gradually increase and decrease producing an increasing and decreasing force on the electrons in the grid wires. Thus a current is induced in the grid at the frequency of the a-c. This current is so small at low frequencies that it may be neglected.

At low frequencies the changes in current through the valve are so gradual that the number of electrons approaching the grid at any instant is practically the same as the number leaving the grid. At very high frequencies, when the period of a cycle is nearly equal to the electron transit time a different effect takes place due to the bunching and spreading of electrons. At low frequencies, due to the uniform distribution of electrons across the valve, while an electron is approaching the grid there is a gradually increasing force of repulsion on electrons in the grid which is cancelled by a gradually decreasing force of repulsion due to a corresponding electron receding from the grid. At very high frequencies, due to electron bunching while an electron is approaching the grid and producing an increasing force of repulsion on the electrons in the grid there may not be an electron receding from the grid to cancel this force. Similarly, while an electron is receding from the grid producing a decreasing force of repulsion there may not be an approaching electron to produce an increasing force. Thus there is a different effect of forces acting upon electrons in the grid wires. As a result, a greater alternating current is caused to flow in the grid circuit at the frequency of the oscillator. Considerable grid current may flow even though the grid is biased negatively with respect to the cathode. This causes a further loss of energy from the circuit and reduces the efficiency. Grid current flow can be reduced by reducing the electron transit time.

Transit time is reduced in two ways.

(i) The electrons are speeded up by using higher anode voltages.

(ii) The distance between cathode and anode is reduced by spacing these electrodes closer together. The breakdown voltage of the valve therefore limits the minimum electron transit time.

5.15 Klystrons. Electron transit time which is a detriment to efficiency in conventional oscillator valves is used to advantage in *klystron* valves. Before proceeding to the detailed description of these valves, it is necessary to explain some new concepts relating to their general operation.

A. *Velocity Modulation*

Fig. 5-34 shows a circuit drawn along conventional lines to show how a stream of electrons may be velocity modulated. The heater, cathode and accelerating grid constitute an electron gun. Electrons which are given off by the cathode are attracted toward the grid by the high positive potential from the supply battery. Most of these electrons pass through the grid wires due to its open construction. The stream of electrons moving with uniform velocity then comes under the influence of the *buncher* grids. These grids are connected to the opposite ends of an oscillatory circuit L_1C_1. The spacing between these grids, the period of oscillation of L_1C_1 and the velocity of the electrons are so adjusted that an electron passes from A to B in the time required for one half oscillation of L_1C_1. If an electron approaches the buncher when A is passing through zero a-c potential its velocity is not changed. By the time it reaches grid B, the potential will again be passing through zero so that its velocity is not affected. An electron which passes through grid A when the grid is positive is accelerated. Half a cycle later it passes through grid B and is further accelerated because grid B is then positive. An electron that passes through the buncher half a cycle later is decelerated by both grids. The stream of electrons leaving the buncher is therefore *velocity modulated.* In passing through the drift space between the buncher and the *catcher,* the accelerated electrons

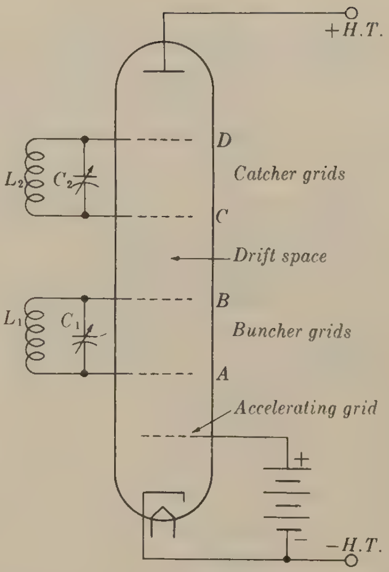


FIG. 5-34 Velocity modulation of the electron stream.

tend to catch up with those ahead while the decelerated electrons slow down to mix with those behind. Consequently the stream is bunched into groups at the point where it enters the catcher circuit.

The catcher is similar in construction to the buncher. It is arranged to absorb energy from the modulated electron stream so that oscillations are reinforced in L_2C_2. If electrons approach grid C when it is negative, the electrons are retarded and energy is absorbed from them. At this instant grid D is positive. Half a cycle later when the electrons approach grid D its polarity has reversed so that energy is again absorbed. If electrons approach grids C and D when they are positive the electrons are accelerated and energy is given up to the electrons. However, the drift space is such that when C and D are negative the approaching electrons are bunched and a great deal of power is absorbed by the catcher circuit. Very little power is given out to the spaced electrons on the other half cycle. The electrons leave the catcher with a lower average velocity and are returned to the battery circuit by the collector plate. In order to maintain oscillations in the whole circuit, some energy from the catcher circuit must be fed back to the buncher circuit at the same frequency and in the proper phase to effect the bunching operation. This is accomplished by a feedback line from the catcher to the buncher circuit. Oscillations are readily started in the high-Q catcher circuit by any non-uniformity of electron emission or operating voltages. The feedback circuit then causes oscillations to increase until sufficient energy is supplied to the catcher to compensate for energy dissipated in the catcher and buncher and also for the energy delivered to an external load.

B. *Resonant Cavities*

The conventional oscillatory circuit consisting of a coil and a capacitor assumes a somewhat different appearance at the frequencies generated by klystron valves. The evolution from LC circuit to *cavity resonator* is shown by the diagrams of Fig. 5-35. The coil and capacitor of (a) are reduced in size to the single-turn inductor and the capacitance between the lead wires shown in (b). In order to reduce losses by radiation from this open circuit, a number of such components placed together make up the structure shown in (c). To reduce losses still further, the ends are closed by changing the shape to that of (d). This doughnut-shaped hollow metal chamber is called a *resonant cavity*. It is electrically a very high-Q oscillatory circuit. When constructed with a diameter of one to three inches, its resonant frequency is in the range 3000 to 10,000 Mc/s.

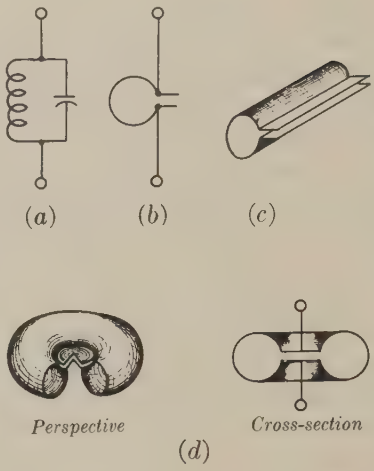


FIG. 5-35 The development of a cavity resonator.

C. *A 3000 Mc/s Klystron*

Two 3000 Mc/s (10 cm. wavelength) klystrons are shown in Fig. 5-36 and Fig. 5-37 respectively. A circuit diagram of the latter valve is shown in Fig. 5-38. A uniform stream of electrons is produced by the heater, cathode, accelerating grid and focusing shield. These electrons are directed through the buncher grids where they become velocity

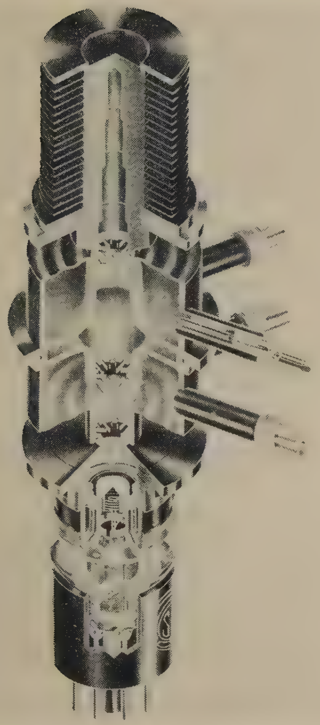


FIG.5-36 A 3000 Mc/s klystron.

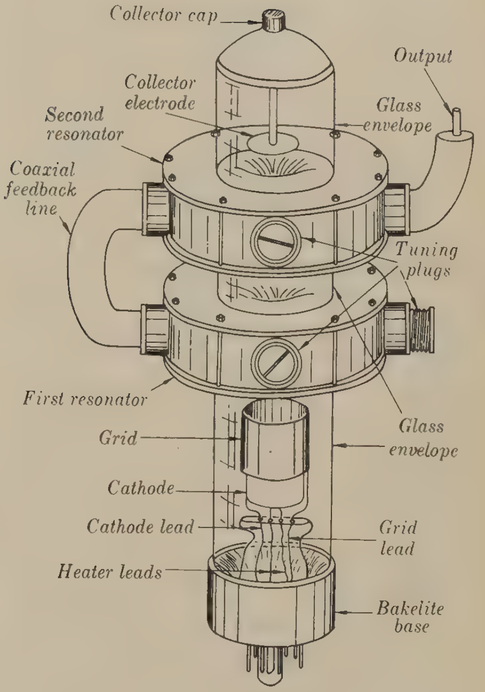


FIG.5-39 Line diagram of a klystron.

modulated at the frequency of the oscillations in this resonant cavity. In the drift space the electrons form into groups just before entering the catcher. The electron groups maintain oscillation in the second resonant cavity. A part of its energy is fed back through the coaxial cable to the buncher cavity to maintain oscillations in it. Output power is taken from the catcher by way of a coaxial

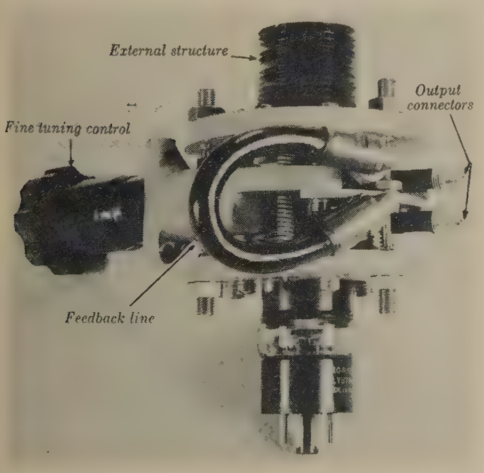


FIG.5-37 A 3000 Mc/s klystron.

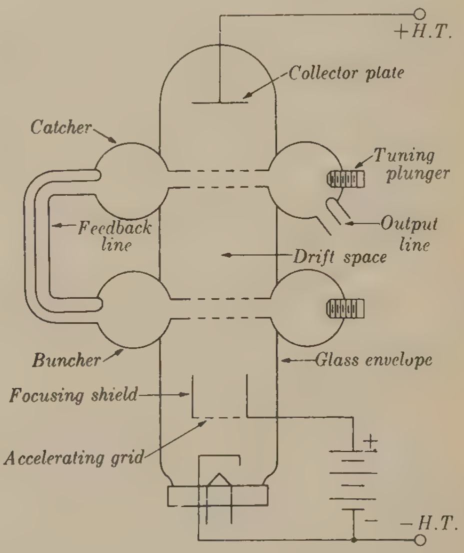


FIG.5-38 Circuit diagram of a klystron.

cable connected to a single turn inductor placed inside the resonant cavity. The strong magnetic field in the cavity induces a current in this loop at the frequency of the oscillations generated.

The klystron valve may be used as an ultra-high-frequency amplifier. Input is applied by way of a coupling loop inserted into the buncher cavity. The feedback is omitted or reduced to prevent oscillation. Output is obtained from the catcher cavity as in the case of the oscillator.

Considerable adjustment of the klystron circuit is required to ensure maximum efficiency. Adjustments are usually provided

(i) to vary the potential on the accelerating grid and hence the speed of the electrons entering the catcher.

(ii) to vary the potential on the collector anode. The velocity of the electron stream must be adjusted so that the electrons pass through the space between the buncher grids in one half the period of oscillation. Also the bunching must reach its maximum as the electrons enter the catcher grids.

(iii) to adjust the frequency of oscillation of both cavities. This is done by plungers which screw into the cavities to alter their resonant frequencies.

(iv) to adjust the amount of feedback and output. This is accomplished by rotating the hairpin coupling loop inside the cavities. Coupling is a maximum when the plane of the loop is perpendicular to the plane of the cavity.

A d-c milliammeter is included in the collector circuit and some method of measuring the r-f output is required for adjustment. The operating potentials and coupling are adjusted to provide maximum r-f output energy for minimum d-c current through the valve. Since these controls are quite interactive correct adjustment of the klystron is very difficult to achieve. A similar oscillator which is less critical in this respect is the *reflex* klystron.

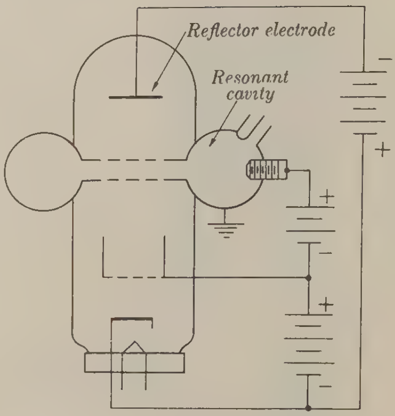


Fig.5-40 Circuit diagram of a reflex klystron.

The electron gun and first cavity resonator of the reflex klystron (Fig. 5-40), are identical in construction and operation with the two-cavity klystron. In place of the second cavity and collector however, the reflex klystron contains a reflector electrode which is operated at a high negative potential. It therefore repels the velocity-modulated stream of electrons in the space between the cavity and the reflector. The bunched electrons are turned back to the cavity in the proper phase to give energy to the resonator. The spent electrons are then removed from the circuit by the cavity itself or by the accelerating grid. The frequency of oscillation is adjusted as before by means of the plungers in the resonant cavity. The potential on the reflector electrode has a secondary control on the frequency as it alters the transit time and hence the phase of the feedback energy. It is adjusted to provide maximum r-f power from the output loop.

A photo of a reflex klystron is shown in Fig. 5-41. Klystrons which are

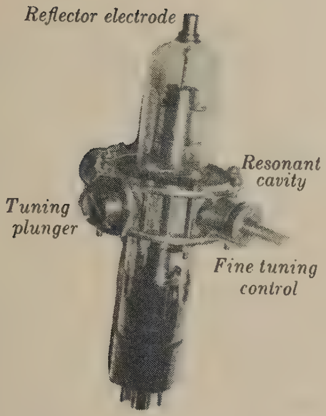


FIG.5-41 Reflex klystron.

designed to operate in the region of 10,000 Mc/s and above are usually tuned by flexing the metal boundaries of the resonant cavity. A slight change in thickness of the cavity causes a considerable shift in its resonant frequency.

The r-f output power of the klystron is very small, usually of the order of a few milliwatts. It is useful as a local oscillator in radar receivers.

5.16 Magnetrons. A *magnetron* is a two-element valve containing a cathode which emits electrons and an anode cylinder which receives the electrons. A high d-c voltage applied between these two electrodes sets up an electric field which tends to make the electrons move radially from the cathode to the anode. In addition to this electric field a strong magnetic field is set up within the valve by a heavy permanent magnet mounted externally. The magnetic field is set at right angles to the direction of electron movement from cathode to anode (Fig. 5-42). The electrons are deflected at right angles to the magnetic field. Therefore an electron follows a curved path in its flight from the cathode to the anode. The curvature of its path depends upon the relative strengths of the electric and magnetic fields. Fig. 5-43 shows three possible paths obtained by keeping the supply voltage constant and increasing the magnetic field strength. In (a) the magnetic deflection is small and the electron path is almost a straight line. In (b) a value of magnetic field strength is reached at which the electron is deflected through 180° on its flight from cathode to anode. It is then on the point of missing the anode completely and being turned back toward the cathode. If the magnetic field strength is increased slightly past this point, the flow of current from cathode to anode ceases abruptly since the electrons are all returned to the cathode. This is illustrated by (c). The value of the magnetic field strength at which this occurs is called the *critical value*. These electron paths are sections of a curve which is called a *cycloid*.

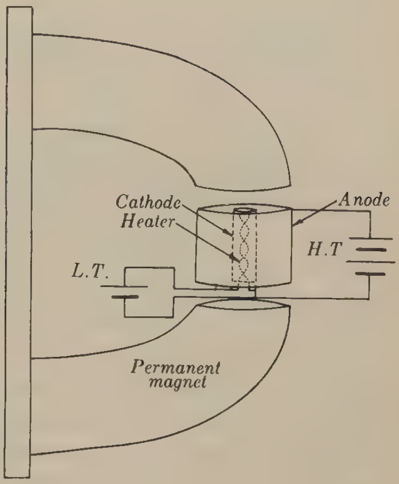


FIG.5-42 A magnetron circuit.

The condition indicated by the curve in (c) is determined by measuring the d-c current through the valve as the

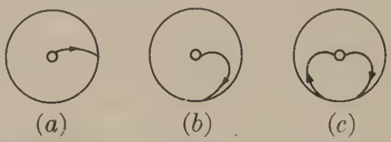

(a) (b) (c)

FIG.5-43 Electron paths with increasing magnetic field.

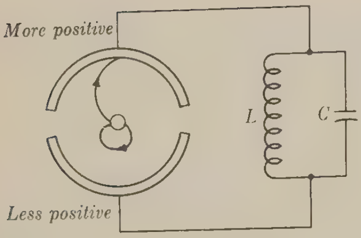


FIG.5-44 A split-anode magnetron showing two electron paths.

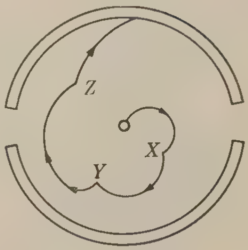


FIG.5-45 Cycloidal electron paths in a split-anode magnetron.

magnetic field strength is adjusted. The anode current cuts off abruptly when the critical field strength is reached. Under these conditions the magnetron is capable of producing oscillations of ultra high frequencies in an external circuit. The frequency is determined by the time taken for the electrons to travel from cathode to anode and return to the cathode. Two types of magnetrons are described in the following sections:

(i) the basic *split-anode* magnetron

(ii) the *multi-cavity* magnetron.

A. *Split-Anode Magnetron*

The split-anode magnetron contains a cylindrical anode, divided into two sections as shown in Fig. 5-44. The two sections are supplied with the same high positive d-c potential relative to the cathode for accelerating electrons from the cathode toward the anode. The anode sections are insulated from each other at the frequency of the oscillations and are connected to the opposite sides of an oscillatory circuit represented by *LC*. The magnetic field is directed into the plane of the diagram and at right angles to it. This causes the electrons to be deflected in a clockwise direction in their flight from cathode to anode. The magnetic and electrostatic fields are adjusted to the critical values described in the preceding paragraph. An alternating voltage across the oscillatory circuit makes one of the anode sections more positive and the other less positive than the critical voltage. If these potentials persisted, an electron leaving the upper side of the cathode with zero or small initial velocity would be drawn to the top anode section while an electron from the bottom of the cathode would be returned to the cathode. The electron paths of two such electrons are shown in Fig. 5-44.

The electron paths are modified however by the changing voltage across *LC*. The electron emitted from the upper side of the cathode follows a path which tends to be cycloidal. This causes the electron path to be somewhat like that shown in Fig. 5-45. It consists of several cycloidal sections of decreasing curvature. At the end of the first section indicated by *X* the electron comes close to striking the cathode but is turned back toward the anode because its velocity has been reduced to the point at which the magnetic field loses control. The electric field re-directs the electron away from the cathode. A similar phase of flight is observed at *Y* and at *Z*. During the next cycloid the electron strikes the anode and its flight is ended. The electron which leaves the opposite side of the cathode, however, follows a path of greater curvature and is returned to the cathode at the end of one cycloid, even when the alternating voltage exists between anode sections.

Energy considerations are of prime importance in explaining the maintenance of oscillations. Fig. 5-46 shows the

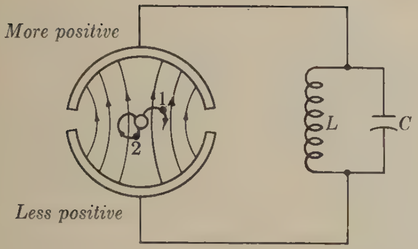


Fig.5-46 Electrostatic field in a split-anode magnetron.

electrostatic lines of force between the two sections of the anode due to the alternating voltage across the *LC* circuit at a particular instant. The d-c component of the electrostatic field may be neglected because it is uniformly distributed from cathode to anode and has similar effects on all electrons in the cathode-anode space. Electron 1 in Fig. 5-46 travels in a direction opposite to the lines of force of the a-c field while crossing the gap between the two anode sections. It is therefore decelerated by the field and gives energy to the oscillatory circuit. Half a cycle later the electron crosses the gap on the opposite side of the cathode. By this time the oscillatory circuit has reversed the field and the electron again travels against the lines of force. It therefore delivers energy to the oscillatory circuit twice during each cycle.

Electron 2, however, travels in the same direction as the lines of force while crossing the gap. It is therefore accelerated and receives energy from the oscillatory circuit. Since electron 1 crosses the anode gap several times while electron 2 is limited to one crossing in its flight, more energy is delivered to the oscillatory circuit than is absorbed from it by the two electrons. These two electrons are representative of the entire number which flow from cathode to anode during normal operation of the magnetron. They are high-efficiency converters of d-c energy from the supply to a-c energy in the oscillatory circuit.

B. Multi-Cavity Magnetron

The oscillatory circuit of the basic magnetron may be replaced for ultra-high-frequency operation by a resonant cavity. The inductance is that of the curved inner surface of the cavity and the capacitance is that across the gap between the anode sections. The anode is divided into several segments containing resonant cavities which are connected to a control chamber. Fig. 5-47 shows the structure of an anode containing eight resonant cavities. Since the cavities cannot be tuned there is no way of varying the frequency of oscillation of the valve once constructed. The electron paths and maintenance of oscillation are basically the same as described for the two-segment magnetron. R-f energy is coupled from the circuit by means of a single-turn loop which extends into one of the resonant cavities. A magnetron used for generating radar transmitter pulses is shown in Fig. 5-48. A cut-away view of this magnetron in its mount is shown in Fig. 5-49.

The average r-f power which a magnetron can produce is not large because of the heat dissipated at the anode by electron bombardment. The anode is usually fitted with cooling fins and its

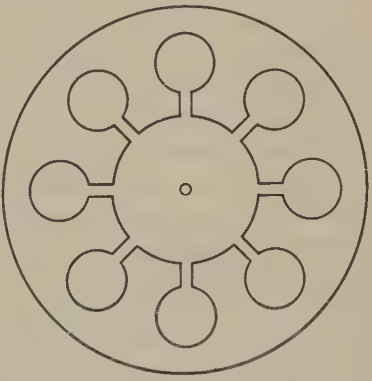


Fig.5-47 An eight-cavity magnetron.

FIG.5-48 Multi-cavity magnetron.

temperature maintained within safe limits by forced air or water cooling. A further limit on the power available is imposed by the cathode which is also subject to considerable electron bombardment by spiralling electrons. Care must therefore be taken that the valve is not operated at excessive voltages which would cause cathode disintegration.

While the average power output is relatively small, the peak power which a magnetron can furnish is extremely large. For this reason it is suitable as the transmitting valve in radar equipments which require pulsed operation. It is common practice to turn on the transmitter for a pulse of one microsecond duration every 1000 microseconds. In this case a peak power output of 100 kilowatts may be maintained for the duration of the pulse with an average output of only 100 watts.

As the anode block is exposed, it is usually operated at d-c ground potential. The valve is turned on by applying a large negative voltage to the cathode for the duration of the pulse. The modulator which operates the magnetron must therefore be capable of delivering a rectangular negative pulse of the order of 10,000 volts at $10A$. The pulse must have a duration of about one microsecond and must be repeated at regular intervals several hundred times per second.

Correct operating potentials are of the utmost importance in the satisfactory operation of the magnetron. Even though the pulse duration is only one microsecond, several thousand oscillations take place during each pulse. It is essential that the frequency of the r-f output remain constant during this time and during each pulse. The optimum operating conditions vary with different magnetron designs and are determined by the manufacturer. They are usually defined by a specific value of applied

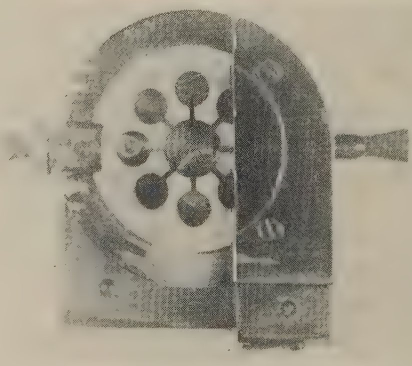


FIG.5-49 Cut-away view of multi-cavity magnetron.

voltage or of average current through the valve.

In some cases the correct operating conditions are determined experimentally each time a new magnetron is put in use. This may be done by monitoring the pulses of current through the valve on an oscilloscope. There are two common methods of obtaining this pulse waveform.

(i) A small resistor is placed in series with the magnetron and the voltage developed across it is applied between the vertical-deflecting plates of the oscilloscope. It shows a square waveform of voltage, which is the outline of the modulation pulse.

(ii) A small amount of the magnetron output is detected to obtain the modulation envelope and this voltage is applied to the oscilloscope.

If the horizontal outline of the pulse is irregular it indicates that the oscillations are unstable. A well-defined split in this line shows that the mode of oscillation changes during the pulse. If the line remains flat but is not sharply defined, the oscillations vary from pulse to pulse. A change in operating potential is required to avoid these conditions.

5.17 Questions and problems.

1. Draw a circuit diagram of a tuned-grid oscillator with series feed. Explain its operation.
2. What advantage does a parallel-feed oscillator have over a series-feed circuit?
3. Draw a tuned-anode oscillator
   (a) with series feed
   (b) with parallel feed.
4. (a) What is meant by the efficiency of an oscillator?
   (b) How is increased efficiency achieved in the modified tuned-anode oscillator?
5. Draw and explain the operation of the circuit of a tuned-anode tuned-grid oscillator.
6. What are the impedances which must be matched in an oscillator circuit?
7. What is meant by the term angle of current flow?
8. Why is it an advantage for an oscillator to have a small angle of current flow? How is this achieved?
9. Why is grid current flow necessary in an oscillator?
10. Why is automatic bias preferable to fixed bias in an oscillator?
11. What is meant by squegging in an oscillator?
12. Explain how the impedance of the valve changes during one cycle of operation. When is it a maximum? When is it a minimum?
13. In what respect are constructional details simpler in a Hartley oscillator than those in the tuned-anode or tuned-grid oscillator?
14. What is the purpose of the by-pass capacitor from the centre tap of the tuned circuit to ground in the series-feed Hartley?
15. Explain why the Hartley circuit must be tuned inductively.
16. What is the purpose of the choke coil between anode and the supply battery in a parallel-feed Hartley oscillator?
17. How does the modified Hartley circuit differ from the conventional Hartley?
18. Why must an oscillatory circuit have a high Q?
19. What factors tend to reduce the Q of an oscillatory circuit?

20. How are these effects prevented from lowering the Q beyond the allowable amount?
21. How does a push-pull oscillator differ from a push-pull amplifier?
22. State three advantages of push-pull oscillators over single-valve types.
23. What is meant by frequency drift?
24. State three chief causes of frequency drift. How are these effects overcome in oscillators used in transmitters?
25. What is meant by an m.o.p.a. transmitter? Explain briefly the function of each of its two sections.
26. How does an electron-coupled oscillator differ from a tuned-anode oscillator?
27. What advantages does an electron-coupled oscillator have over those types employing triode valves?
28. Explain what is meant by the piezoelectric effect.
29. Draw the equivalent circuit of a crystal. How does its Q compare with that of LC circuits at radio frequencies?
30. Which of the inductance, capacitance and resistance values is responsible for giving the crystal its very high Q?
31. What is meant by frequency temperature coefficient of a crystal? In what unit is it measured?
32. What shape of crystal has a very low temperature coefficient?
33. What is meant by a negative temperature coefficient?
34. Make a diagram of the Miller crystal oscillator. How is its LC circuit tuned?
35. Which of the LC tuned-circuit oscillators is similar in circuit design to the tri-tet oscillator?
36. Explain how the Pierce oscillator is the crystal equivalent of the Colpitts circuit.
37. What limits the power output of a Pierce oscillator?
38. What is the lower frequency limit of crystal oscillators?
39. What is the upper frequency limit of crystal oscillators?
40. How may this limit be extended by the following stages?
41. What factors reduce the Q of LC circuits designed for very-high-frequency oscillators?
42. How are these factors reduced by the use of transmission lines as oscillatory circuits?
43. How is the frequency of a transmission-line oscillator related to the length of the transmission lines?
44. What two types of transmission lines are used in oscillator circuits?
45. Draw a diagram of a push-pull oscillator that employs transmission lines in both the grid and anode circuits.
46. How is feedback provided in transmission-line oscillators?
47. What is the purpose of using transmission lines in cathode circuits?
48. Compare the length and grounding of the cathode lines
    (a) when the cathode is to be at high r-f potential
    (b) when the cathode is to be at ground r-f potential.
49. How are the heater lines designed to prevent r-f feedback to the filament transformer?
50. State one advantage which a concentric line has over an open transmission line in an oscillator circuit.
51. State one disadvantage of the concentric-line oscillatory circuit.

52. How does interelectrode capacitance limit the frequency of an r-f oscillator?

53. What constructional details are employed to reduce interelectrode capacitance?

54. What is meant by electron transit time?

55. Why is electron transit time not an important consideration at low radio frequencies?

56. How does electron transit time reduce the efficiency of ultra-high-frequency valves?

57. How may electron transit time be reduced?

58. In what respect is there conflict between reduction of transit time and reduction of interelectrode capacitance?

59. Why does considerable a-c current flow in grid wires at high frequency, even though the grid is biased negatively with respect to the cathode?

60. Why is this induced grid current more pronounced at high than low radio frequencies?

61. How is electron transit time used to advantage in a klystron valve?

62. Explain what is meant by velocity modulation of electrons.

63. What is the purpose of the drift space in a klystron?

64. Why can the catcher absorb more energy from the bunched electrons than the buncher loses in modulating them?

65. What is the purpose of the resonant cavity connected to a klystron?

66. How does the *Q* of a resonant cavity compare with that of an *LC* circuit having the same resonant frequency?

67. What factors contribute to this high value of *Q*?

68. How is energy coupled into and out of a resonant cavity?

69. How is the frequency of a resonant cavity adjusted?

70. Draw a circuit diagram of the reflex klystron.

71. What is the purpose of
    (a) the electron gun
    (b) the resonant cavity
    (c) the reflector electrode in a reflex klystron?

72. What use is made of klystron valves in radar equipment?

73. What use is made of magnetron valves in radar sets?

74. What is the purpose of the magnet in a magnetron?

75. Draw a cross-sectional diagram of a split-anode magnetron.

76. What is the purpose of the resonant cavities in this valve?

77. Describe the path of an electron from cathode to anode in the magnetron.

78. How is the output energy obtained?

79. What makes the magnetron a suitable transmitting valve in a radar set?

80. What are the requirements of a modulator supplying the magnetron?

81. What test is made to ensure stable oscillation of the magnetron?

82. What factor has the greatest effect on the stability of oscillation?

CHAPTER VI

AMPLITUDE MODULATION

6.1 Principle of modulation. Audio intelligence is transformed into an audio-frequency current by a microphone when the intelligence is in the form of speech or by a keyed audio oscillator when the intelligence is in the form of a coded message. This audio-frequency current which is called the *signal* may be transmitted by wire over great distances. However, it can not be transmitted over an appreciable distance through space with an efficiency high enough to be practical. Thus, transmission through space must take place at radio frequencies. The audio-frequency intelligence is superimposed upon a high-frequency radio wave called a *carrier* by a process called *modulation*.

6.2 A simple method of modulation. Fig. 6-1 shows a simple modulated-oscillator circuit consisting of a parallel-feed tuned-anode oscillator with a microphone M coupled to its grid circuit. L_1 and C_1 form the tuned anode circuit, C_2 and L_3 the parallel feed and R and C the automatic series-type grid-leak bias. The feedback is provided by the coupling between L_1 and L_2. The microphone M is coupled to the grid circuit of the oscillator by the iron-core a-f transformer T.

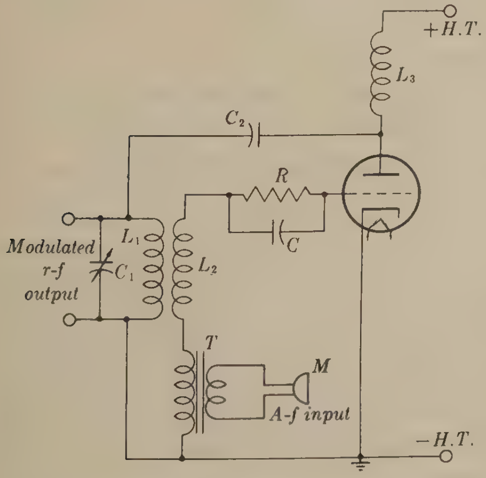


FIG.6-1 A simple modulated r-f oscillator.

When no sound waves strike the microphone there is no a-f voltage across the secondary of the transformer T. Hence the output from the oscillator, taken across the tuned anode circuit L_1C_1 is an r-f voltage of constant amplitude, Fig. 6-2 (a), and is said to be *unmodulated*. The bias provided by R and C adjusts itself to such a value that only a small pulse of anode current flows during the positive peak of each cycle of the r-f to maintain oscillation in the tuned anode circuit L_1C_1.

When sound waves strike the microphone, however, an a-f voltage is developed across the secondary of T, Fig. 6-2 (b). On the positive half cycles of the a-f voltage the negative bias on the valve, already provided by R and C, is decreased. On the negative half cycles the negative bias is increased.

On the first quarter cycle of the a-f, while the bias decreases, larger r-f pulses of current flow through the valve, increasing the amplitude of the oscillating current in L_1C_1 and hence increasing the amplitude of the r-f voltage output across L_1C_1 to a maximum value, Fig. 6-2 (c). On the second quarter cycle the bias goes more negative, returning to its original value and the amplitude of the r-f output voltage decreases to its original unmodulated amplitude. On the third quarter cycle the bias con-

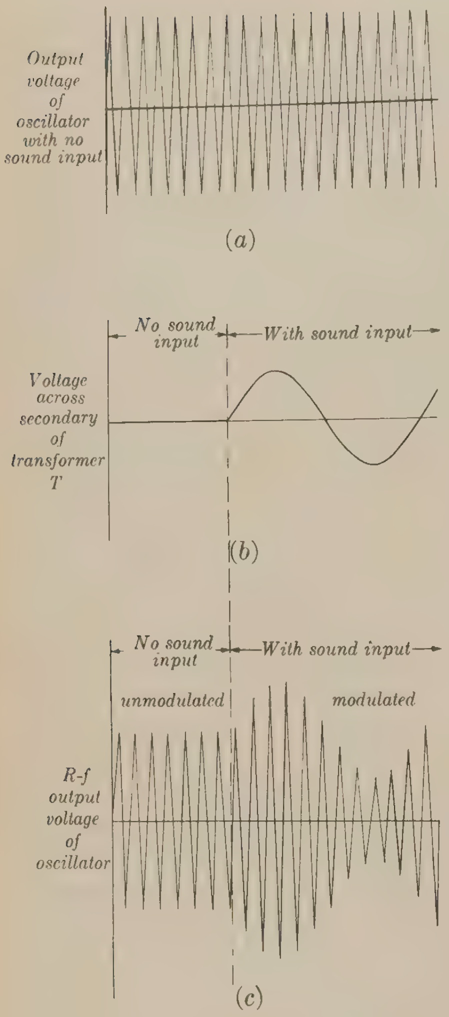

Output voltage of oscillator with no sound input
(a)
No sound input
With sound input
Voltage across secondary of transformer T
(b)
No sound input
With sound input
unmodulated
modulated
R-f output voltage of oscillator
(c)

Fig. 6-2 Voltage waveforms in a simple modulator (a) Unmodulated r-f output of oscillator (b) Modulating a-f input to oscillator (c) Modulated r-f output of oscillator.

tinues to go more negative and the amplitude of the r-f voltage continues to decrease to a minimum value. On the fourth quarter cycle the bias decreases to its original value and the amplitude of the r-f output voltage increases to its original amplitude. That is, as the bias is caused to vary at an a-f rate the amplitude of the r-f output voltage of the oscillator varies at the same a-f rate. The r-f output which is varying in amplitude is called an *amplitude-modulated* r-f.

If the intensity of the sound input to the microphone is increased, the variation in bias is greater and hence the variation in amplitude of the modulated r-f is greater.

The r-f carrier may be modulated at any stage in the transmitter circuit. The modulating voltage is applied usually to an r-f power amplifier stage. The r-f oscillator itself can be modulated but this causes the oscillator frequency (carrier frequency) to be unstable.

6.3 Degree of modulation. A single-frequency audio modulating wave is illustrated in Fig. 6-3 (a). This sine wave has an amplitude A and may represent either current or voltage. An r-f carrier wave of amplitude B is illustrated in Fig. 6-3 (b). When the carrier is modulated by the audio wave the resulting waveform appears as in Fig. 6-3 (c). The outline of the modulated r-f wave is called the *modulation envelope.* The audible output at the receiver for a given carrier depends upon the *depth of modulation A.* As a result the largest possible variations in A are usually employed. When the amplitudes A and B are equal, the amplitude of the envelope is reduced to zero every cycle of the a-f signal, Fig. 6-3 (d).

The degree of modulation is usually expressed as *percentage modulation.*

$$\textit{Percentage modulation} = \frac{A}{B} \times 100$$

where A = amplitude of the modulation envelope
B = average envelope amplitude.

In Fig. 6-3 (c), $A = \frac{1}{2}B$ and the percentage modulation is 50%. In Fig. 6-3 (d), $A = B$ and the percentage modulation is 100%. This wave is also described as a *completely modulated* wave.

The degree of modulation is sometimes expressed as a *modulation factor m.*

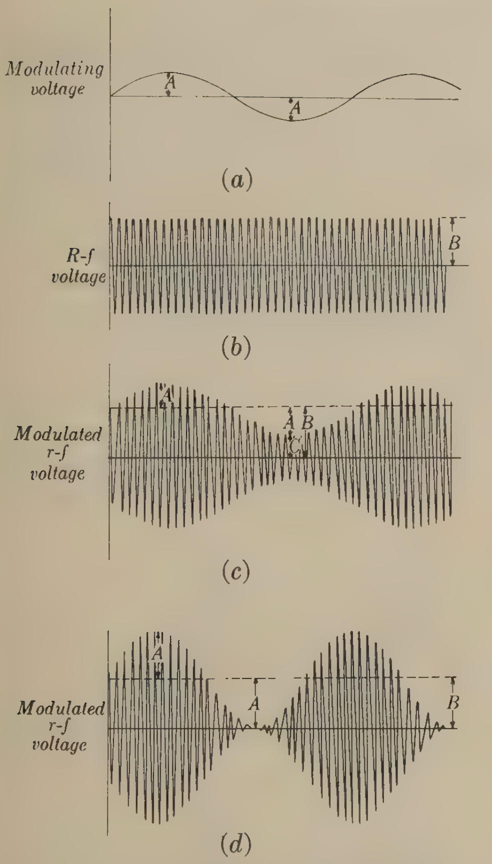


FIG. 6-3 Degree of modulation of r-f carrier, (a) Modulating a-f voltage (b) Unmodulated r-f carrier voltage (c) R-f carrier modulated 50% (d) R-f carrier modulated 100%.

That is,

$$m = \frac{B - C}{B}$$

where m = modulation factor
B = average envelope amplitude
C = minimum envelope amplitude.

The degree of modulation varies from a modulation factor of 1 with 100% modulation to a modulation factor of 0 with no modulation.

If the carrier is modulated more than 100%, a waveform as in Fig. 6-4 results. There is an interval in which the transmission is cut off. The modulated wave is distorted since the modulation envelope does not have the same waveform as the original modulating sine wave.

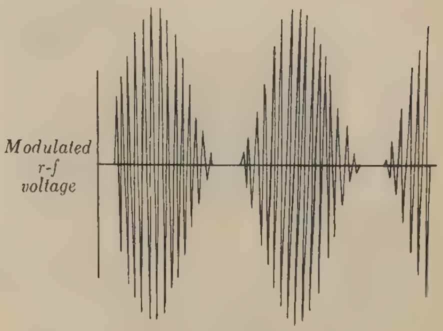


FIG. 6-4 R-f voltage modulated more than 100%.

This produces a distorted sound at the receiver output since the audio wave in the receiver varies in accordance with the distorted modulation envelope. Since there is distortion, the audio output contains harmonics of the original modulating audio frequency.

Distortion may result with less than 100% modulation if the modulation envelope and modulation signal differ in waveform. However, if the modulation envelope has exactly the same waveform as the modulating signal, as in Fig. 6-3 (a) and (c), then the signal is transmitted without distortion.

6.4 Methods of amplitude modulation. In the circuit of Fig. 6-1 the a-f modulating voltage is introduced into the grid circuit. However, the modulation may also be accomplished by introducing the modulating voltage into the anode or cathode circuits. As a result there are three principal methods of amplitude modulation.

(i) Anode modulation
(ii) Grid modulation
(iii) Cathode modulation.

6.5 Anode modulation. A typical circuit for anode modulation is shown in Fig. 6-5. The r-f carrier input, as shown in Fig. 6-6 (a), is applied to the grid of the r-f power amplifier which is operated Class *C*. This r-f input comes from another r-f power amplifier called the *driver*, which in turn receives its input

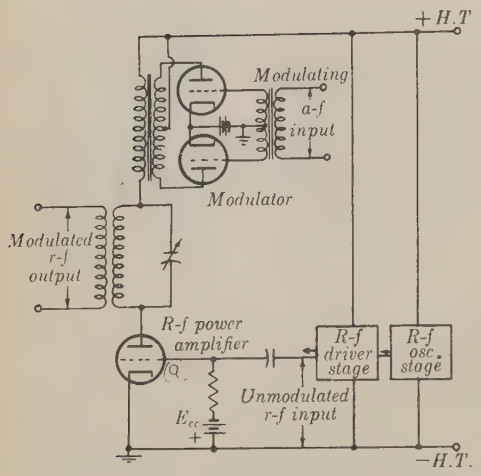


FIG.6-5 Anode modulation circuit.

from the r-f oscillator. The tank circuit in the anode circuit of the power amplifier is tuned to the frequency of the r-f carrier. The modulated output from the power amplifier is taken from the coil of the tank circuit by transformer coupling.

The a-f modulating voltage, which originates in such a device as a microphone or a record player pickup does not have sufficient amplitude to produce the desired degree of modulation. Hence this a-f voltage is applied to a-f power amplifiers, the final output of which is applied to the r-f amplifier. The a-f power amplifier or amplifiers form the *modulator*.

The modulator in the circuit of Fig. 6-5 is a push-pull a-f power amplifier. Fig. 6-6 (b) shows a single-frequency a-f voltage present in the output from the modulator. The modulator is transformer coupled into the anode circuit of the r-f power amplifier and hence this type of modulation is called *anode modulation.*

A. Modulation Process

For a given amplitude of r-f input voltage to the r-f amplifier circuit, the amplitude of the r-f output voltage from the circuit is zero if the H.T. voltage is zero. If the H.T. voltage is gradually increased from zero, the steady component of anode voltage also increases and the amplitude of the r-f output from the circuit increases. Similarly a gradual

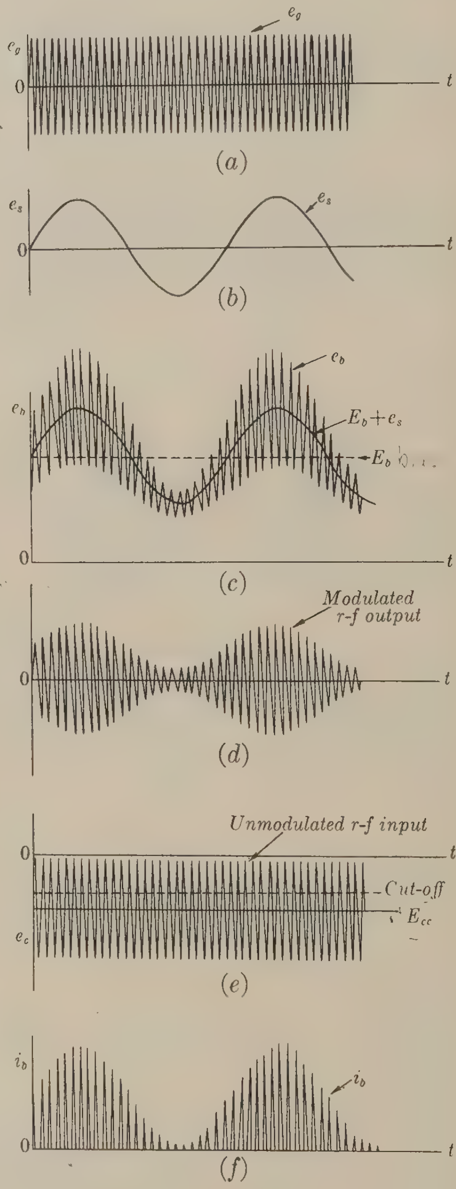


FIG.6-6 Voltage waveforms in anode modulation (a) R-f carrier input to grid circuit (b) A-f modulating signal to p.a. anode circuit (c) Anode voltage of p.a. (d) Modulated r-f output voltage of p.a. (e) Grid-to-cathode voltage of p.a. (f) P.a. anode current.

decrease in H.T. voltage and resulting decrease in the steady component of anode voltage causes a decrease in r-f output.

The power amplifier of Fig. 6-5 has two controlling voltages on its anode. There is the steady anode voltage E_b supplied by the H.T. supply and superimposed upon this is the a-f modulating voltage e_s. Thus the resultant controlling anode voltage, which in an ordinary r-f amplifier is the steady component of anode voltage, is varying at an audio rate as illustrated by the curve E_b+e_s in Fig. 6-6 (c). The r-f carrier applied to the control grid has a constant amplitude. As the anode voltage rises at the audio rate, the r-f voltage output from the power amplifier gradually increases and as the anode voltage falls at the audio rate the r-f voltage output gradually decreases. Thus, the output from the power amplifier is an r-f voltage whose amplitude varies at the audio rate. The modulated r-f voltage e_b at the anode is shown in Fig. 6-6 (c) superimposed upon the power amplifier anode voltage which is varying at the audio rate.

Although audio-frequency currents are present in the tank circuit of the r-f power amplifier, the modulated output voltage contains no a-f voltage since the tank circuit has negligible impedance for a-f. The modulated r-f output which contains no d-c or a-f components is illustrated in Fig. 6-6 (d).

Since the r-f power amplifier is operated Class C, the steady negative grid bias E_{cc} is beyond cut-off, Fig. 6-6 (e). The r-f input varying about the steady bias causes the grid voltage to rise above cut-off only at the peaks of the positive half cycles. Hence the anode current flows in pulses, Fig. 6-6 (f).

B. Conditions for No Distortion

The r-f power amplifier must be designed so that the least possible distortion appears in the modulation envelope. For no distortion the envelope must vary in accordance with the audio modulating signal. For no distortion with anode modulation, there must be a linear relationship between the anode current and the anode voltage varying at the audio rate. This is illustrated in Fig. 6-7 for a Class C amplifier which has an r-f carrier voltage applied to the grid. As the steady anode voltage varies at the audio rate, the pulses of r-f anode current have a modulation envelope which varies exactly with the modulating signal.

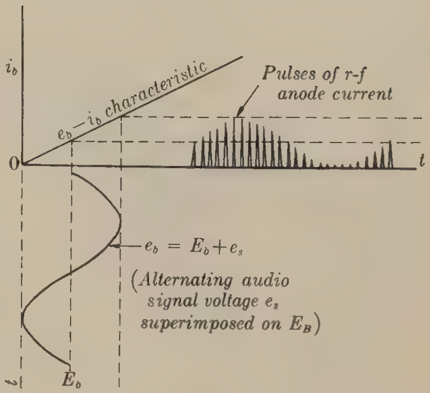

i_b
$e_b - i_b$ characteristic
Pulses of r-f anode current
0
t
$e_b = E_b + e_s$
(Alternating audio signal voltage e_s superimposed on E_B)
E_b
t

FIG. 6-7 Conditions for distortionless anode modulation.

Also, for the least possible distortion to appear in the modulation envelope, there must be a linear relationship between the anode tank circulating current i_o, which provides the modulated r-f output from the tank, and the anode voltage varying at the audio rate. This is illustrated in Fig. 6-8. Instead of flowing in pulses, the tank circuit current has the waveform illustrated due to the oscillatory action in this circuit.

C. Percentage Modulation

The output from the r-f power amplifier can be modulated 100% by causing the anode voltage to vary from zero to twice the value of the steady anode supply voltage when the a-f modulating voltage e_s is applied. Fig. 6-6 (c) illustrates anode voltage variations for less than 100% modulation. The anode voltage variations for 100% modulation are such that the modulating voltage e_s on one half cycle reduces the anode voltage from E_b to zero, Fig. 6-9 (a). At this instant of zero anode voltage, there is zero anode current, Fig. 6-9 (b), and the r-f output from the amplifier is zero, Fig. 6-9 (c). On the other half cycle of modulating voltage, e_s raises the anode voltage from E_b to $2E_b$ and at this instant the r-f output from the amplifier is a maximum. To obtain these large anode voltage variations, the a-f power output of the modulator must be large. For this reason anode modulation is also known as *high-level modulation.* If the power of the modulating signal must be kept low, it is necessary to modulate an early r-f stage in the transmitter where the power involved is small.

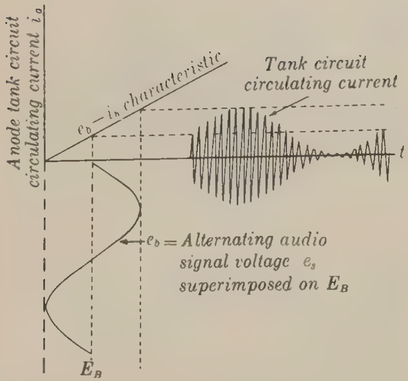
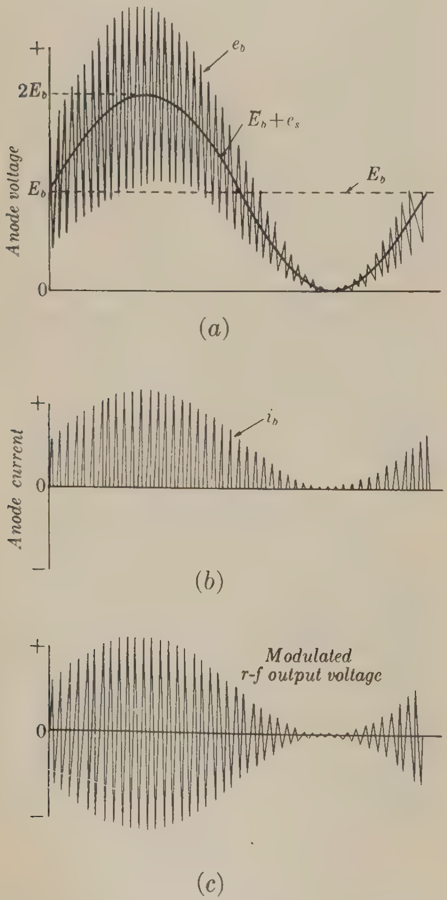


FIG. 6-8 Graph of anode voltage vs circulating current in tank circuit in distortionless anode modulator.



(a)

(b)

(c)

FIG. 6-9 Voltage and current waveforms in r-f amplifier for 100% modulation (a) Anode voltage (b) Anode current (c) Output voltage.

D. Other Circuits

Fig. 6-10 illustrates another circuit arrangement for anode modulation. The modulator and the r-f power amplifier are in parallel. The a-f input to the modulator results in an amplified a-f output across the a-f modulator choke L_1 in the modulator anode circuit. Corresponding audio changes in the anode voltage of the r-f amplifier take place and anode modulation results. R drops the d-c voltage applied to the r-f power amplifier anode so that this voltage is lower than that on the modulator anode. This permits very high percentage modulation to be attained. C is an a-f by-pass capacitor. The r-f choke L keeps r-f out of the modulator circuit.

Screen-grid valves of the pentode and beam power type may be anode modulated providing the modulation is applied to anode and screen grid at the same time in order that the modulating voltage have sufficient control over the current.

6.6 Grid modulation. An r-f power amplifier circuit designed for control-grid modulation is shown in Fig. 6-11. The power amplifier is operated Class *C* with the grid biased beyond cut-off by E_{cc}. The r-f output from the driver is coupled through C_c to the grid of the power amplifier. This r-f input to the power amplifier is the r-f carrier of steady amplitude. The a-f modulating voltage is coupled to the grid of the amplifier through T_1.

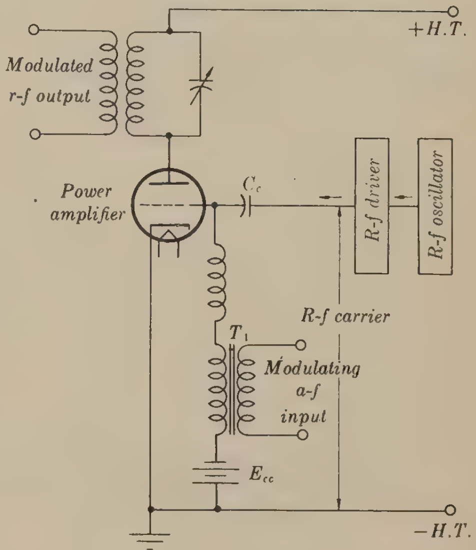


FIG. 6-11 Grid modulation circuit.

A. Modulation Process

The operation of the power amplifier depends on the fact that for a given amplitude of input to the circuit, the amplitude of the output r-f voltage is increased with decreasing grid bias and is decreased with increasing grid bias. The grid bias is caused to increase and decrease at the audio rate by the modulating signal which is applied to the grid circuit. As a result, the amplitude of the r-f output voltage varies at the audio rate and the required modulated r-f voltage is produced.

Fig. 6-12 (a) shows the r-f carrier voltage applied to the power amplifier control-grid circuit and Fig. 6-12 (b) shows the a-f modulating signal also applied to the control-grid circuit. Fig. 6-12 (c) shows the following three voltages on the r-f power amplifier control grid: the steady grid bias E_{cc}, the modulating audio signal e_s which causes the bias to vary at the audio rate and the r-f carrier input e_c. Fig. 6-12 (d) shows the corresponding pulses of anode current resulting from the circuit being biased beyond cut-off. Since the output tank circuit oscillates, the output voltage has the appearance of a sine wave, which varies in amplitude at the audio rate. Fig. 6-12 (e) shows the resulting modulated r-f output, the modulation envelope varying in accordance with the modulating audio signal.

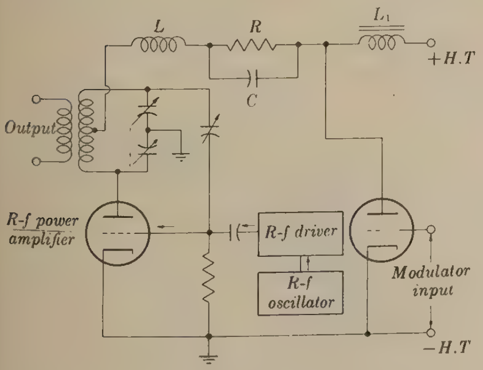


FIG. 6-10 Anode modulation circuit with modulator and r-f amplifier in parallel.

B. Conditions For No Distortion

When the circuit is designed for minimum distortion, the power output is low. A compromise between power output and distortionless modulation is necessary. To obtain large pulses of anode current and hence high power output with minimum distortion the grid voltage is usually driven to zero

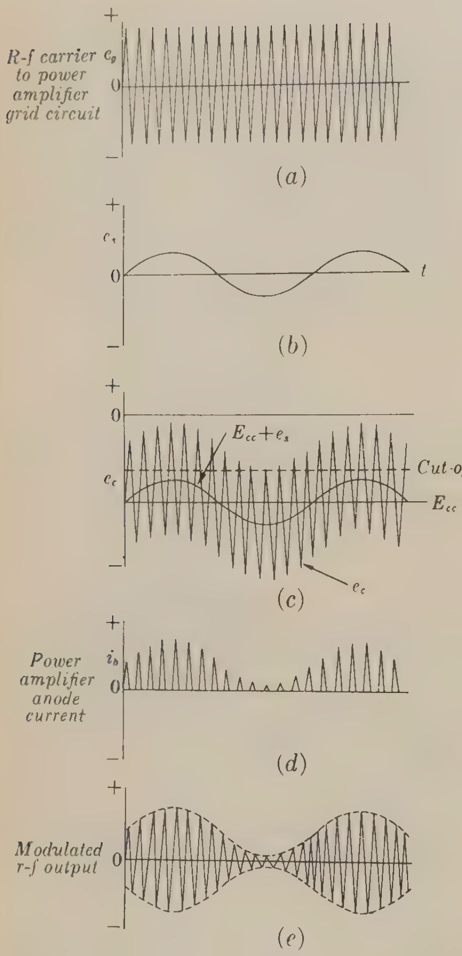


FIG. 6-12 Voltage and current waveforms in grid modulation (a) Carrier input to grid circuit of r-f amplifier (b) A-f modulating input to grid circuit of r-f amplifier (c) Grid-to-cathode voltage of r-f amplifier (d) Anode current of amplifier (e) Modulated r-f output of amplifier.

by the r-f at the peak positive swings of the modulating input. If the grid is driven positive by the r-f on the modulating peaks, the power output increases, but since grid current flows distortion results.

A dynamic $e_c - i_b$ curve which is linear over the operating range is necessary for minimum distortion. When this curve is linear the modulation envelope has the same waveform as the modulating audio signal. The dynamic characteristic required is illustrated in Fig. 6-13. The steady grid bias E_{cc}, the modulating audio signal e_s and the r-f carrier input e_c are shown along with the resulting pulses of anode current.

The advantage of grid modulation is that very little audio-frequency power is required for 100% modulation. For this reason, grid modulation is sometimes called *low-level modulation.* However, the anode efficiency of the r-f power amplifier with grid modulation is lower than with anode modulation. This is due to the lower average value of grid excitation. That is, with anode modulation each positive half cycle of the r-f input may drive the control grid to zero, while with grid modulation the control grid is driven to zero only during the positive peak of the a-f modulating voltage.

6.7 Cathode modulation. *Cathode modulation* is a combination of anode modulation and grid modulation. The circuit, illustrated in Fig. 6-14, has the modulation transformer between cathode and ground. The transformer secondary has two sections L_1 and L_2. The upper end of L_2 is connected to the control grid through the a-f coupling capacitor C and the r-f choke L_3. Since C and L_3 have negligible impedance for a-f, the tapping point of the transformer has the same a-f potential as the grid. The

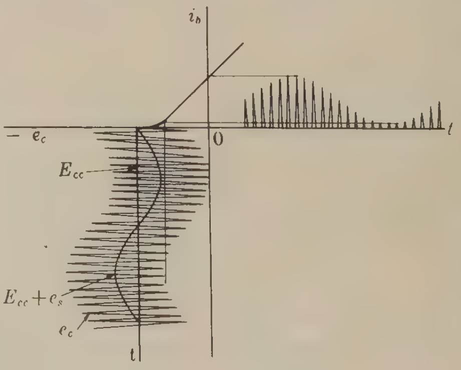


FIG. 6-13 Conditions for distortionless grid modulation.

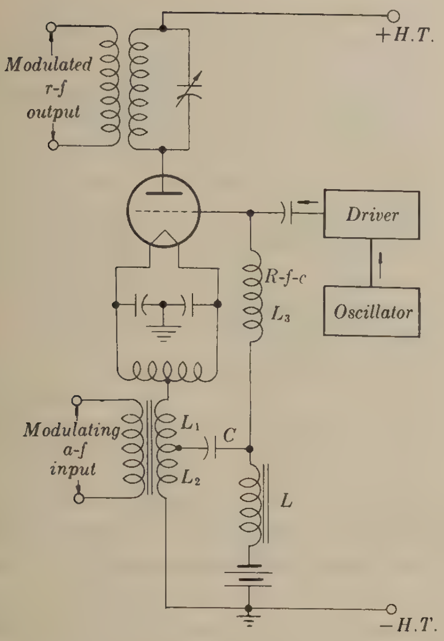


FIG. 6-14 Cathode modulation circuit.

a-f choke L prevents L_2 from being shorted for a-f through the bias battery. The a-f voltage developed across L_2 by the a-f modulating input causes both the grid and cathode potentials to vary by the same amount with respect to ground. This is equivalent to applying a modulating a-f voltage at the anode and thus anode modulation takes place.

The a-f voltage across L_1 causes the cathode potential to vary with respect to the grid potential and thus provides grid modulation.

The anode efficiency of the r-f power amplifier with cathode modulation is between the efficiencies for grid and anode modulation.

6.8 Suppressor-grid modulation. A circuit designed for suppressor-grid modulation is shown in Fig. 6-15. It consists of a parallel-feed Class C pentode r-f amplifier with tuned input and output circuits. The screen supply voltage E_{cc2} is filtered for r-f by C_1, L_1 and C_2. The suppressor grid negatively biased by E_{cc3}, is maintained at r-f ground potential by the filter L_2 and C_3.

The operation of the circuit depends upon the following facts.

(i) When an r-f carrier voltage is applied to the input an amplified r-f voltage appears at the output.

(ii) When the potential of the suppressor grid is made more negative the amplitude of the r-f output voltage is decreased.

(iii) When the potential of the suppressor is made less negative the amplitude of the r-f output voltage is increased.

Thus, if a steady bias E_{cc3} is applied to the suppressor to set the operating point and then an audio modulating signal is applied to the suppressor via the transformer T_1, the suppressor bias is caused to vary at an audio rate and the amplitude of the r-f output, for steady r-f carrier input, varies at the audio rate. Thus the modulated r-f voltage is obtained at the output of the amplifier.

To obtain the value of steady operating bias for the suppressor, a negative bias is applied to the suppressor grid and increased negatively until the r-f output from the circuit becomes zero

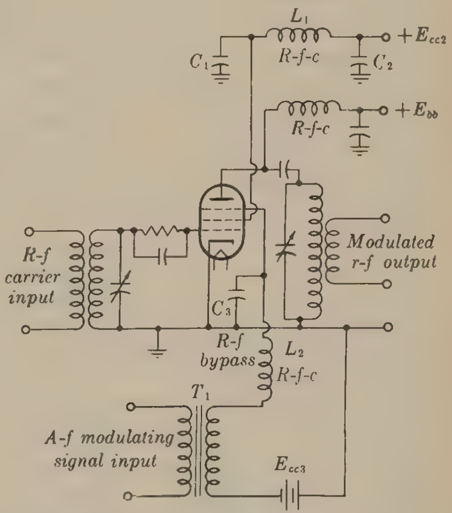


FIG. 6-15 Suppressor-grid modulation circuit.

when the r-f carrier input is applied. For example, in Fig. 6-16 which shows typical operating curves, a suppressor bias −150V is required. This value of negative suppressor grid voltage is noted. Then the suppressor bias is gradually decreased, increasing the r-f output and the bias is noted at the point where the r-f output begins to drop off. This is approximately +50V in Fig. 6-16. The value half way between these two bias limits is the operating bias. This is about −50V in Fig. 6-16. The maximum amplitude of the a-f modulating voltage which may be used is half the difference between the two bias limits. This would be 100V for Fig. 6-16.

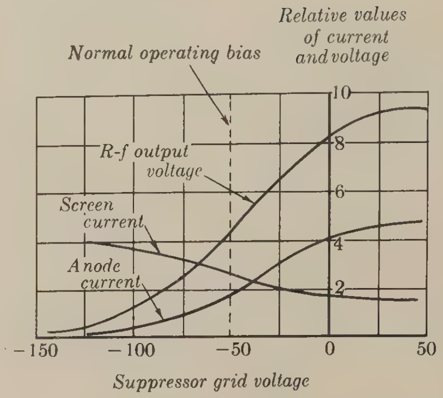


Fig. 6-16 Operating curves for suppressor-grid modulation.

The r-f output curve in Fig. 6-16 is not particularly linear except near the operating point. Thus, to avoid distortion of the modulation envelope, the amplitude of audio modulating voltage input and hence the percentage modulation is restricted. Also, since the screen current rises as the suppressor potential goes negative the modulating voltage amplitude must not be such as to cause the permissible screen dissipation to be exceeded.

6.9 Comparison of amplitude modulation methods. A comparison of the two common methods of modulation, grid modulation and anode modulation, indicates that anode modulation is fairly free of distortion, is efficient and requires large audio power. Grid modulation is less efficient and has more distortion but the modulating power required is small. Code transmitters, where some distortion is permissible, use this method.

A comparison of the various methods of modulation is shown in the table at the bottom of this page.

6.10 Sidebands. When an r-f carrier is modulated by a single sinusoidal audio frequency, there are three sinusoidal radio-frequencies in the modulated carrier. They are:

(i) the original carrier frequency
(ii) the carrier frequency plus the audio frequency
(iii) the carrier frequency minus the audio frequency.

To illustrate graphically that three r-f frequencies are present in the single-frequency modulated wave, the development of a 100% modulated r-f carrier

Comparison of Methods of Modulation

Modulation Type	P.A. Anode Efficiency	Output Power Relative to Valve Size	A-F Power Required	Relative Distortion	Where Used
Anode	high	high	high	low	r-t transmitters
Grid	medium	medium	low	medium	code transmission
Cathode	medium-high	medium	medium	medium	code transmission
Suppressor	medium	medium	low	medium	code transmission

is shown in Fig. 6-17. The r-f carrier current or voltage wave of frequency 100 Kc/s and amplitude A_0 is sinusoidal, Fig. 6-17 (b). Fig. 6-17 (a) illustrates a second sinusoidal r-f wave of frequency 90 Kc/s and amplitude $\frac{1}{2}A_0$. A third sinusoidal r-f wave of frequency 110 Kc/s and amplitude $\frac{1}{2}A_0$ is shown in Fig. 6-17 (c).

If the instantaneous values of the three r-f waves are added algebraically at corresponding times and the resultant values plotted on a graph, the resultant wave obtained appears as in Fig. 6-17 (d). This resultant wave is a modulated r-f wave having a frequency of 100 Kc/s but modulated at a frequency of 10,000 c/s. This modulation frequency is a-f and in value is the difference between the r-f carrier frequency of 100 Kc/s and either of the second or third r-f waves of 90 Kc/s and 110 Kc/s.

This development shows that a single-frequency audio-modulated r-f wave is made up of three sinusoidal r-f components. The two r-f frequencies 90 Kc/s and 110 Kc/s on either side of the carrier are called the *sideband frequencies.* The frequency which is greater than the carrier frequency is called the *upper sideband frequency* and that which is smaller is called the *lower sideband frequency*. Fig. 6-17 also shows that the amplitudes of the two sideband frequencies are each one-half the amplitude of the carrier for 100% modulation. Since the r-f carrier, modulated at an audio rate, is made up of three r-f frequencies, it is important to note that it contains no audio-frequency components.

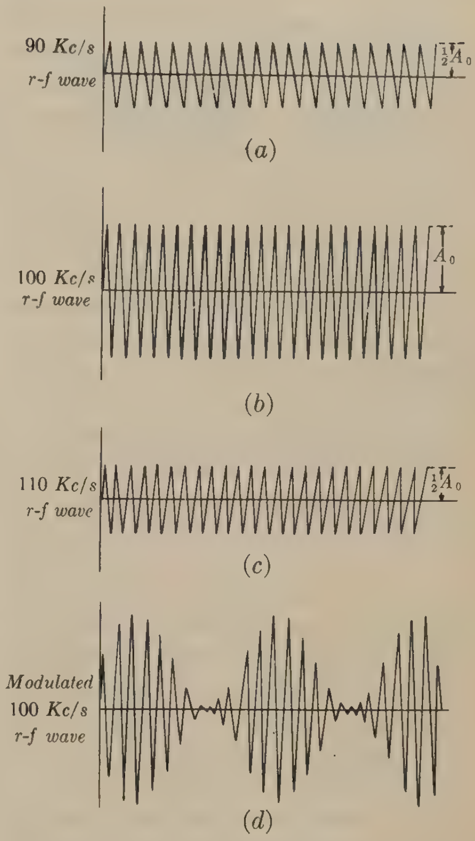


FIG.6-17 Frequency components in modulated r-f waveform (a) Lower sideband r-f waveform (b) R-f carrier waveform (c) Upper sideband r-f waveform (d) Resultant modulated r-f carrier waveform.

When an r-f carrier is modulated by the frequencies of voice or music, many audio frequencies become involved. For each audio frequency in the modulation process, two new r-f frequencies are present in the modulated r-f wave in addition to the original r-f carrier frequency f. They are:

(i) the carrier frequency plus the audio frequency $(f+f_s)$

(ii) the carrier frequency minus the audio frequency $(f-f_s)$.

The many new r-f frequencies resulting, form a group or band of frequencies called *sidebands*. Therefore it is important to note that in amplifiers of modulated r-f the response curves of the tank circuits must be wide enough so that their impedances are high for all the sideband frequencies present in the modulated wave.

A modulated r-f carrier in a long-wave or short-wave radio band occupies a group of radio frequencies which extend about 5000 cycles on each side of the carrier, when voice or music frequencies are involved. This group of frequencies form the *channel.* The channel width is 10,000 cycles or twice the highest modulation frequency ($2f_s$).

From Fig. 6-17 it is seen that for 100% modulation, the ratios of the amplitudes of the carrier and two sideband currents or voltages are A_0: $\frac{1}{2}A_0$: $\frac{1}{2}A_0$, that is, 1: $\frac{1}{2}$: $\frac{1}{2}$. Since power is proportional to the square of the current or voltage, the ratios of the powers of the carrier and two sidebands are 1:$\frac{1}{4}$:$\frac{1}{4}$. Thus,

$$\frac{\text{total power of carrier and sidebands}}{\text{power of carrier}} = \frac{1+\frac{1}{4}+\frac{1}{4}}{1} = \frac{1.5}{1}.$$

From this it may be seen that the continuous power with 100% modulation is 1.5 times the power of the original unmodulated carrier. Further, with 100% modulation, 66$\frac{2}{3}$% of the power is in the carrier and 33$\frac{1}{3}$% is in the sideband frequencies, that is, the sideband power is one-half the carrier power. Since the intelligence is conveyed by the sideband frequencies it is advisable to keep the sideband power as high as possible.

The power required to generate the carrier wave is supplied from the power supply of the r-f power amplifier. However, the power required to generate the sideband components of the modulated wave is supplied from the modulator output. The following example illustrates how the preceding information may be used.

Example. A 5000-watt carrier is to be modulated 100%. Find the required modulator power output if the r-f amplifier has an anode efficiency of 60%.

The total power of carrier and sidebands = 1$\frac{1}{2}$×5000 = 7500 watts. Therefore, the power of the sidebands = 7500 − 5000 = 2500 watts. Since the r-f power amplifier has an anode efficiency of only 60%, the modulator must supply $\frac{100}{60}\times 2500$ or approximately 4200 watts.

6.11 Types of transmission with amplitude modulation. In the types of amplitude modulation previously discussed the a-f modulating signal is a single-frequency sinusoidal voltage. In actual practice the r-f carrier may be modulated in any one of the following ways.

A. Radio-Telephony Modulation (r-t)

In this type of transmission the amplitude of the continuous r-f carrier is varied in accordance with voice or music frequencies (Fig. 6-18). In the case of voice or music, many audio frequencies are involved, ranging up to about 15,000 c/s.

B. Keyed Continuous-Wave (Keyed c-w)

If the intelligence is in the form of Morse code, the transmission is keyed. The radiated wave is started and stopped to correspond with the dots and dashes of the code. During the periods of transmission when the key is closed, the amplitude and frequency of the r-f wave are constant. During the intervals when the key is open, the amplitude is zero and there is no transmission. The waveform of the resultant transmission appears in Fig. 6-19.

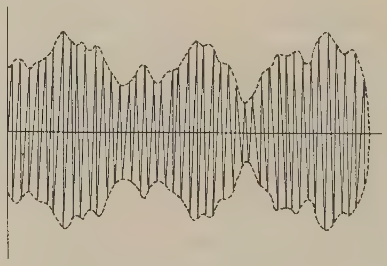


FIG.6-18 R-t modulated r-f carrier waveform.

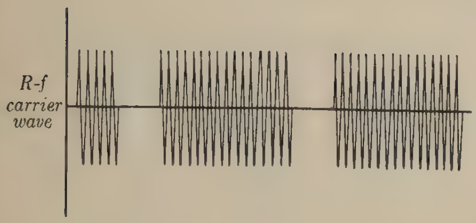


FIG.6-19 Keyed c-w waveform.

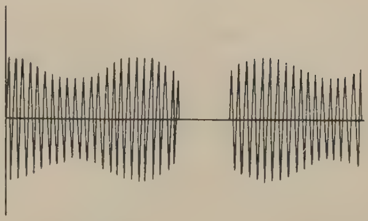


FIG.6-20 Keyed m-c-w waveform.

C. *Keyed Modulated-Continuous-Wave (Keyed m-c-w)*

This type of transmission is the same as keyed *c-w*, except that the amplitude of the wave to be keyed is varied at a single audio frequency. Typical a-f's are 400 and 1000 c/s. Fig. 6-20 illustrates *keyed m-c-w* or keyed *single-frequency modulation.*

6.12 Keying principles. The type of transmitter required for keyed c-w transmission (Fig. 6-19) is similar to that designed for r-t modulation except that there is no modulated r-f amplifier and no modulator. For keyed *m-c-w* transmission (Fig. 6-20) these stages are present. The amplifiers are operated Class *C* at the fundamental frequencies or are operated as frequency multipliers depending upon the operating frequency.

Crystal control is commonly used for short-wave code transmitters. For frequencies above 10 Mc/s frequency-multiplier amplifiers are usually employed, the crystal-controlled oscillator working at a sub-multiple of the operating frequency. Long-wave code transmitters usually have master-oscillator control without a crystal.

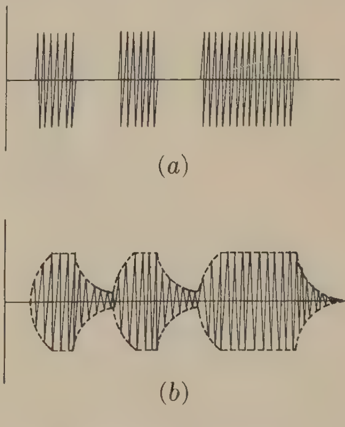
The following problems may occur in keying and may be overcome as suggested.

A. *Back-Wave*

The transmitter is keyed satisfactorily if the power output from the transmitter and aerial is zero when the key is open and if full power is sent out from the aerial when the key is closed. If there is transmission from the aerial in the intervals between the dots and dashes of the code, the dots and dashes are only louder parts of a continuous sound and the reading of the code is difficult. The sound heard between the dots and dashes is called *back-wave* and might exist, for example, if only the final stage preceding the aerial were keyed. The keying of a high-power transmitter is usually carried out in one of the low-power amplifiers or frequency-multiplier amplifiers.

B. *R-F Pulse Shape*

The shape of the r-f pulses forming the dots and dashes is important. If the transmission is started and stopped instantaneously, the pulses will have the square shape shown in Fig. 6-21 (a). In this case, an infinite number of



(a)

(b)

FIG.6-21 Harmonics generated by keying (a) Sharp keying introduces high-frequency harmonics (b) Distortion of waveform due to suppression of high-frequency harmonics.

harmonics is present. Although the amount of energy involved at any one radio frequency is small, interference may be caused in nearby receivers in the form of clicks or thumps. In order to reduce interference, the higher harmonics in these pulses are suppressed. Lower harmonics are retained to keep dots and dashes clearly separated and the message easily readable. The higher harmonics may be suppressed and the pulses given a suitable shape in either of the following ways.

(i) The r-f energy is passed through tuned circuits with the desired band width and shielding is used to prevent the undesired components from being radiated before passing through the tuned circuits.

(ii) A time-delay or lag circuit is used with the key as in Fig. 6-22. This circuit causes the pulses to be turned on and off gradually so that the very high harmonics are eliminated. The inductance L in the lag circuit causes oscillations to build up slowly and the capacitor C assists in causing them to die away slowly. The circuit is carefully adjusted to avoid extreme distortion of the pulses which causes them to blur into each other, as in Fig. 6-21 (b). The variable resistor R across the inductance and the resistor R_1 in series with the capacitor permit a wide range of adjustment. The optimum values of resistance, capacitance and inductance are found experimentally.

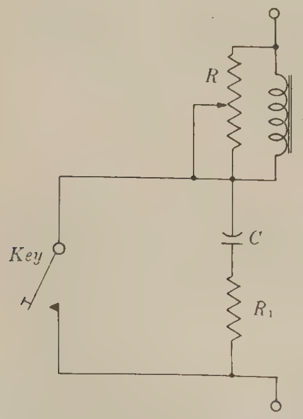


Fig.6-22 A time-delay or lag circuit.

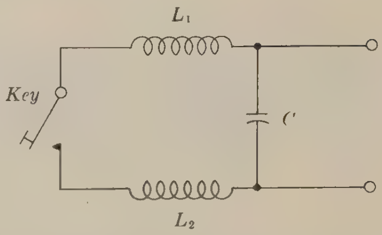


Fig.6-23 Inductance-capacitance spark suppression circuit.

C. *Key Contact Sparking*

A spark at the key contacts, even though small, causes damped oscillation in the keying circuit which may modulate the transmitter output or may simply be radiated by the key wiring. This interference is usually confined to the vicinity of the transmitter. The sparking is reduced or eliminated by a circuit arrangement as in Fig. 6-23. This filter circuit consisting of L_1, L_2 and C is connected across the key and close to it. The chokes, 2-80mH, oppose changes in current and the capacitor, 0.001-0.1μF, absorbs charge to reduce sparking.

6.13 Keying Circuits. The position selected for the key in the transmitter circuit depends on

(i) the power of the transmitter
(ii) the speed of keying
(iii) the degree of frequency stability required.

The two general methods of keying are accomplished

(i) by interrupting the main power supply to the transmitter or to one or more of its r-f stages
(ii) by controlling the grid potential of one or more of the r-f stages.

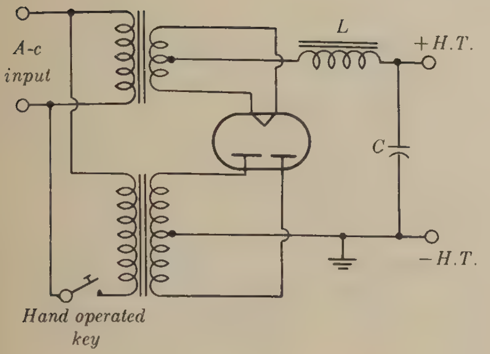


FIG.6-24 Keying in the primary circuit of power supply.

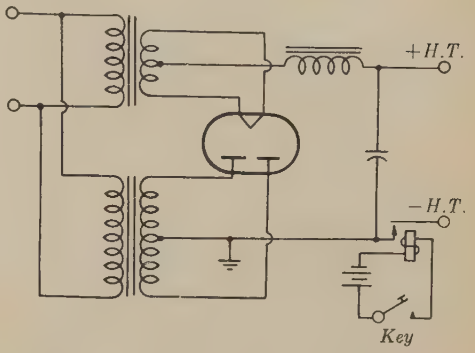


FIG.6-25 Keying in the secondary circuit of power supply.

A. Power-Supply Keying

Power-supply keying is accomplished by either *primary* keying or *secondary* keying.

1. PRIMARY KEYING.

The hand key or relay-operated magnetic key is connected in the primary circuit of the main H.T. supply transformer (Fig. 6-24). The power supply shown in this diagram employs full-wave rectification. Since all stages receiving H.T. voltages from this supply are shut down when the key is open, minimum power is wasted. However, since a large current is being interrupted the key must have heavy contacts and the resulting mechanical inertia prevents high-speed keying. The filter circuit L and C of the power supply causes a keying lag due to the time constant involving the filter capacitor. If more than a single section of filtering is used the time lag is too great for crisp keying.

2. SECONDARY KEYING.

In secondary keying, the key, usually a relay-operated magnetic key, is connected into the secondary of the main H.T. supply transformer either in the positive or negative lead. Since the positive lead is not a good choice due to its high positive potential the key is usually in the negative lead (Fig. 6-25). As in primary keying, all stages fed by this power supply are shut down when the key is open and minimum power is wasted. However, the key contacts are heavy and high-speed keying is difficult.

A variation of this method results when the power supply makes use of grid-controlled rectifier valves (Fig. 6-26). A relay-operated key is employed. Keying is accomplished by applying a suitable bias to the grids to shut off the power supply when the key is open. When the key is closed the keying relay operates to remove the bias and turn on the supply. The resistor R prevents the bias supply from being short-circuited

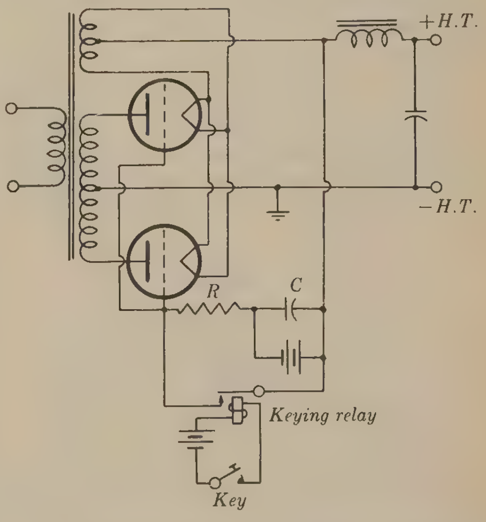


FIG.6-26 Keying in the secondary circuit of power supply using grid-controlled rectifier valves.

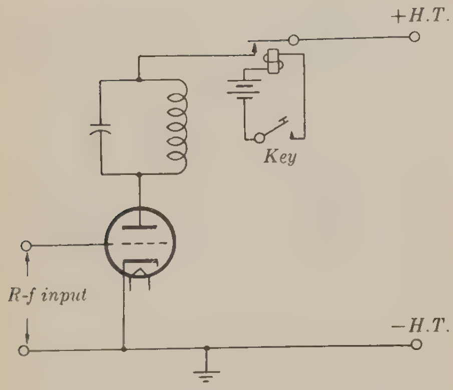


Fig.6-27 Keying in the anode supply lead.

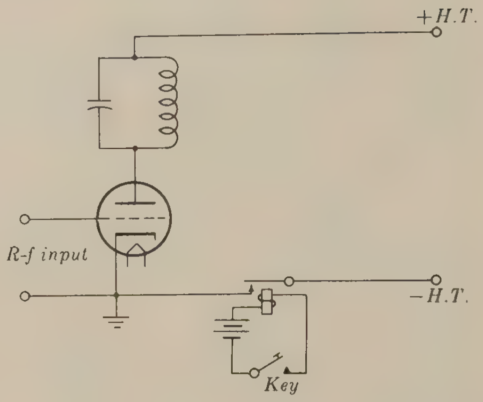


Fig.6-28 Keying in the cathode supply lead.

when the key is down. Sparking across the relay contacts is prevented by C and R. This circuit is used with high-voltage supplies and the keying relay must be well insulated.

3. ANODE KEYING.

Another variation of secondary keying is *anode keying*. The relay-operated magnetic key is inserted in the positive lead as in Fig. 6-27 or the negative lead as in Fig. 6-28.

The method of inserting the key in the positive lead is avoided whenever possible due to the high potential. This type of keying is best suited for low-power work.

4. CATHODE KEYING.

Cathode keying is similar to anode keying since it cuts off the power supply to one stage. Since the keying is in the cathode lead (Fig. 6-29) the grid return path to the cathode is also broken. Less sparking occurs than in anode keying. When the cathode is directly heated this type of keying is called *centre-tap keying* (Fig. 6-30). Since centre-tap keying produces extremely bad key clicks a key click filter consisting of L_1, R_1, C_1, and R_2 is employed.

A more efficient key click filter employs a valve (Fig. 6-31). The keying valve V_1 is connected in series with the centre-tap lead of the keyed valve V_2. When the key is closed the grid of the keying valve is connected directly to the filament. This valve V_1 then acts as a low resistance in the lead. When the key is open a large resistance is placed in the cathode circuit of V_1. V_1 operates near cut-off and is a high resistance in the cathode

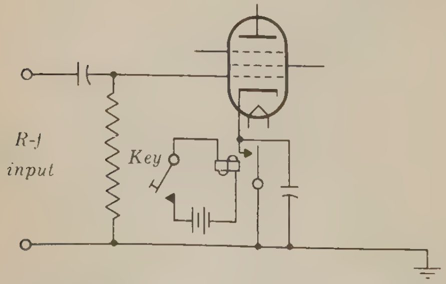


Fig.6-29 Cathode keying with indirectly-heated valve.

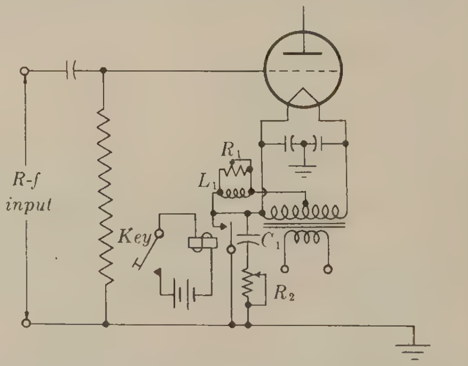


Fig.6-30 Cathode keying with directly-heated valve.

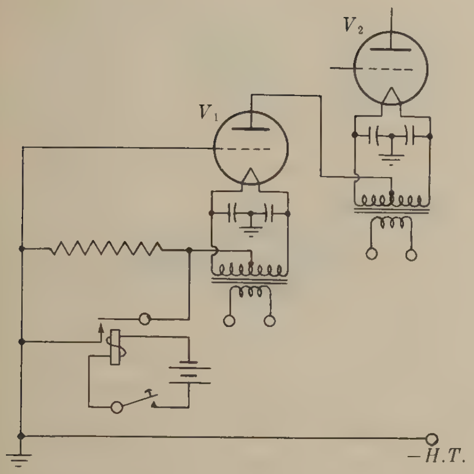


FIG.6-31 Centre-tap keying with a keying valve.

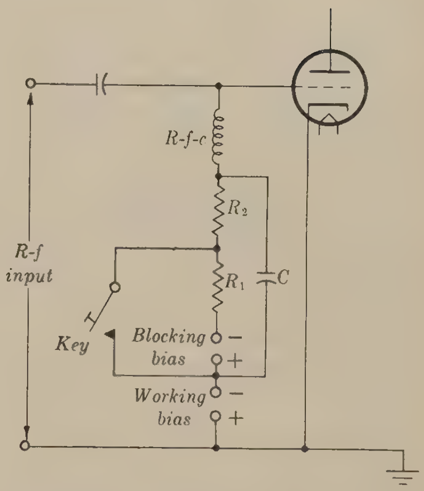


FIG.6-32 Grid keying with blocking bias from a separate source.

circuit of V_2. As a result the output of the amplifier V_2 is reduced to nearly zero.

B. *Grid Keying*

This type of keying, illustrated in Figs. 6-32 and 6-33, is carried out by supplying sufficient *blocking bias* or *paralyzing bias* to a control grid in addition to the normal working bias on the control grid to cut off anode current when the key is open. When the key is closed this blocking bias is removed leaving only the normal working bias.

Fig. 6-32 shows the blocking bias connected in series with the working bias between grid and cathode. Resistor R_1 prevents shorting of the blocking bias when the key is down. R_2 and C form a circuit for controlling the shape of the code pulses.

In Fig. 6-33 the cathode resistor R provides the working bias while the blocking bias is obtained from a voltage divider R_1 and R_2 across the H.T. supply. In some cases both the blocking bias and the working bias are obtained from a voltage divider across the H.T. supply.

Grid keying is always used in high-speed transmission. A small key is employed since only small power is controlled directly. There is minimum risk of sparking.

Absorption grid keying is illustrated by Fig. 6-34. The stage with V_1 is being keyed. V_2 is the *absorber valve* which, with its load resistance R, constitutes a load on the power supply equal to that of the transmitter when the transmitter is shut down during keying. Thus there is the same load on the power supply at all times and there is little possibility of power line surges with the

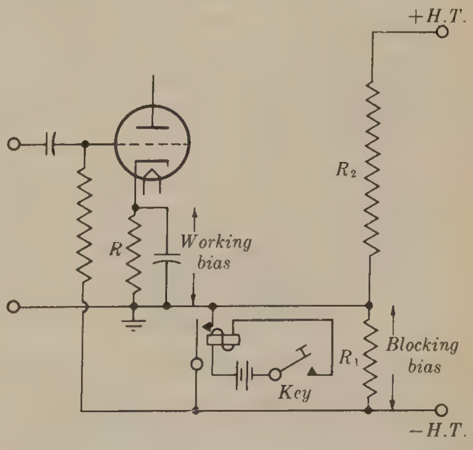


FIG.6-33 Grid keying with blocking bias from the power supply.

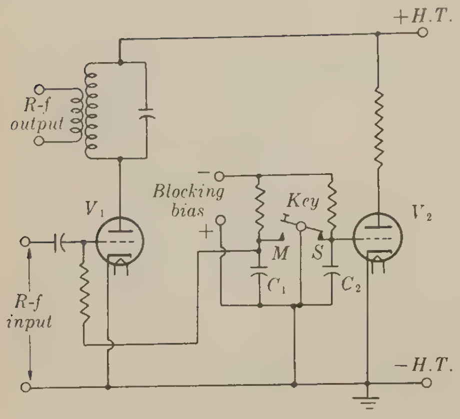


Fig. 6-34 Absorption grid keying.

result that the frequency is held more constant.

During the intervals between code pulses, the key is at S. The grid of the absorber valve is connected to its cathode so that current flows through V_2 and power is dissipated in the load resistor R. At the same time the blocking bias is applied to the keyed stage V_1 to shut it down.

During the transmission of code pulses, the key is at M. The blocking bias is applied to the absorber valve V_2 and the keyed stage has only its operating bias. Capacitors C_1 and C_2 are connected across the key contacts to prevent sparking.

6.14 Oscilloscope test patterns. The oscilloscope test for percentage modulation can be relied upon as an effective method since the oscilloscope gives an actual picture of the modulated r-f output of the transmitter.

Two procedures are employed. The first gives a picture of the modulated r-f wave clearly showing the envelope. The second provides a trapezoidal pattern.

A. First Procedure

To employ the first method the output from the final stage of the transmitter is applied to the vertical-deflection plates of the oscilloscope by means of a pick-up loop and twisted pair line, Fig. 6-35 (a), or link coupling and a circuit tuned to the transmitter frequency, Fig. 6-35 (b). The coupling is adjusted to give the desired amplitude on the oscilloscope screen. The internal sweep of the oscilloscope is adjusted to an audio frequency to give the picture of the modulated r-f output. When there is no modulated r-f output applied to the oscilloscope the sweep alone is observed as in Fig. 6-36 (a). When the carrier only is applied the pattern, Fig. 6-36 (b), is seen. Less than 100% modulation provides a picture as in Fig. 6-36 (c). The pattern in Fig. 6-36 (d) appears with 100% modulation. Over 100% modulation appears as in Fig. 6-36 (e). If the modulation is sinusoidal the percentage modulation is calculated by direct measurement on the oscilloscope screen and the use of the formula:

$$\textit{Percentage modulation} = \frac{h_1 - h_2}{h_2} \times 100$$

where h_1 = the maximum height with modulation

h_2 = the height of the carrier alone.

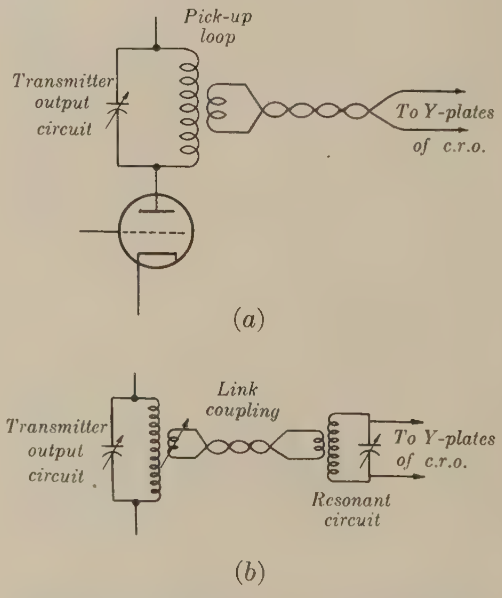


Fig. 6-35 Coupling c.r.o. to transmitter to obtain modulation test patterns.

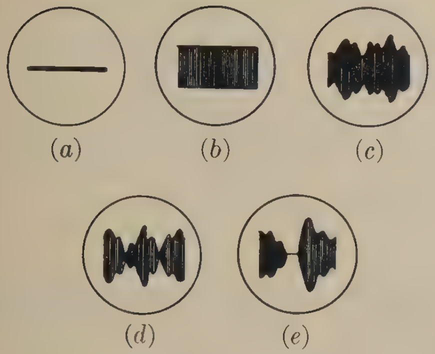

(*a*) (*b*) (*c*) (*d*) (*e*)

FIG.6-36 Modulation test patterns (a) Zero r-f output (b) Unmodulated r-f output (c) R-t modulated r-f output with less than 100% modulation (d) R-t modulated r-f output with 100% modulation (e) R-t modulated r-f output with more than 100% modulation.

B. Second Procedure

The trapezoidal pattern is obtained by applying the transmitter output to the vertical plates of the oscilloscope as previously suggested. However, in this case the internal sweep voltage on the horizontal plates is replaced by a portion of the output from the modulator as in Fig. 6-37. The output voltage of the modulator is applied through the blocking capacitor C to the voltage divider R_1 and R_2. The voltage applied to the X-plates is taken across the centre tap of the potentiometer R_2 and ground. The oscilloscope pattern appears as in Fig. 6-38 (a) when there is no r-f voltage applied to the vertical plates and no modulator voltage applied to the horizontal plates. When there is no modulator output and the carrier is applied to the vertical plates the pattern of Fig. 6-38 (b) is observed. When the carrier is modulated less than 100% and voltages are applied to vertical- and horizontal-plates a trapezium is seen, Fig, 6-38 (c). With 100% modulation the pattern appears as a pointed wedge, Fig. 6-38 (d). Over 100% modulation gives the pattern, Fig. 6-38 (e), in which the wedge has a thin horizontal line appearing at the point of the wedge. The modulation percentage is obtained as in the first method by measuring the modulated and unmodulated carrier heights and using the formula previously stated.

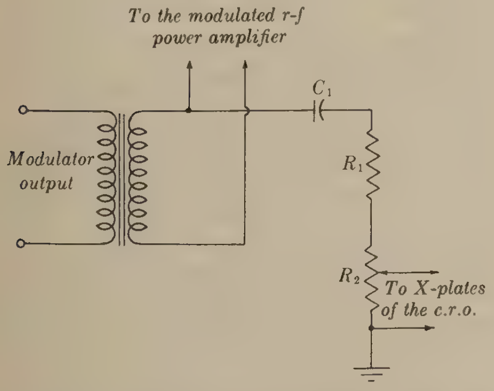

To the modulated r-f power amplifier
C_1
Modulator output
R_1
R_2
To X-plates of the c.r.o.

FIG.6-37 Coupling c.r.o. to transmitter to obtain trapezoidal test patterns

6.15 Types of modulation. In this chapter amplitude modulation only is discussed. However, in general, modulation is the process of producing a radio wave some characteristic of which varies in accordance with the audio modulating wave.

A radio wave has three characteristics which can be varied when modulation is taking place. These characteristics are the amplitude, frequency and phase. As a result, there are three types of modulation, as follows:

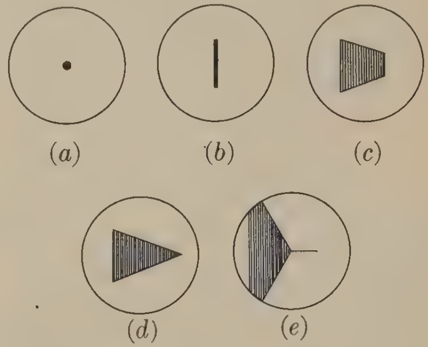

(*a*) (*b*) (*c*) (*d*) (*e*)

FIG.6-38 Trapezoidal test patterns (a) Zero r-f output and zero sweep voltage (b) Unmodulated r-f output (c) R-f output modulated less than 100% (d) R-f output modulated 100% (e) R-f output modulated more than 100%.

(i) *amplitude modulation,* in which the amplitude is varied in accordance with the modulating wave, while frequency and phase remain constant

(ii) *frequency modulation,* in which the frequency is varied while amplitude and phase remain constant

(iii) *phase modulation,* in which the phase is varied while amplitude and frequency remain constant.

The last two types of modulation are discussed in Chapter XIII.

6.16 Applied mathematics.

A. Mathematical Expression of an Amplitude-Modulated Wave

A sinusoidal carrier wave of current may be represented by the formula

$$i = I \sin (\omega t + \theta) \qquad (1)$$

where i = instantaneous r-f current in amperes

I = amplitude of the r-f current in amperes

ω = angular frequency in radians per sec. ($\omega = 2\pi f$)

f = frequency of the r-f current in cycles per second

t = time in seconds

θ = phase angle in radians.

If this carrier is amplitude modulated by a single-frequency sine wave of current its amplitude is given by

$$I = I_0 (1 + m \sin \omega_s t) \qquad (2)$$

where I_0 = average amplitude of modulated current in amperes

m = modulation factor

$\omega_s = 2\pi f_s$ where f_s is the modulating frequency.

Substituting for I from (2) in (1) and assuming $\theta = 0$,

$$\begin{aligned} i &= I_0 (1 + m \sin 2\pi f_s t) \sin 2\pi f t \\ &= I_0 (\sin 2\pi f t + m \sin 2\pi f_s t \sin 2\pi f t) \\ &= I_0 [\sin 2\pi f t + \frac{m}{2} \cos 2\pi\ (f - f_s) t \\ &\quad - \frac{m}{2} \cos 2\pi\ (f + f_s) t]. \end{aligned}$$

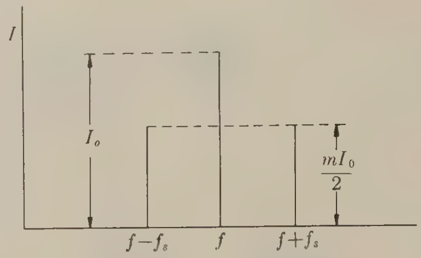


FIG. 6-39 Comparison of amplitudes of carrier and sidebands for single sinusoidal modulating frequency.

This formula for the modulated r-f carrier shows that the wave contains the three frequencies

(i) the original carrier frequency f

(ii) the carrier plus the modulating frequency $(f + f_s)$

(iii) the carrier minus the modulating frequency $(f - f_s)$.

The carrier and the two sidebands are represented graphically in Fig. 6-39, the current amplitude vertically and the frequency horizontally.

When the carrier is 100% modulated, $m = 1$ and the amplitude of each sideband is one-half the amplitude of the carrier.

B. Calculation of C_1, R_1 and R_2 in Fig. 6-37

To calculate the sizes of R_1 and R_2 in Fig. 6-37 the following considerations are made. The ratio of R_2 to R_1 should be large enough to give a satisfactory horizontal sweep voltage on the screen. Consider R_2 to be a $0.25 M\Omega$ potentiometer. The value of R_1 depends upon the audio output voltage of the modulator in the following manner. The total resistance of R_1 and R_2 in series should be $0.25 M\Omega$ for every $150 V$ of modulator output. In general, this voltage is equal to $\sqrt{PR}$ where P is the audio power output of the modulator and R is its output impedance. However, in grid modulation with a 1:1 output transformer it may be assumed that this voltage is $0.7E_b$ for one valve or $1.4E_b$ for a push-pull stage, where E_b is the d-c component of modulator anode voltage. If the transformer ratio is not 1:1, this output voltage is calculated as above and then multiplied by the turns ratio. The blocking capacitor C is $0.1\mu F$ or more. The following example illustrates a calculation of R_1.

Example. Suppose that the d-c component of anode voltage in a grid-modulated push-pull stage is $536V$ and that the output transformer ratio is 1:1. Calculate R_1 if R_2 is $0.25M\Omega$.

$$\text{Output voltage} = 1.4E_b = 1.4\times 536 = 750\,V$$

Since $R_1+R_2 = 0.25M\Omega$ for every $150V$ of output voltage

$$R_1+R_2 = 0.25\times 10^6\times\frac{750}{150} = 1.25\times 10^6$$

$$\therefore R_1+R_2 = 1.25 \text{ megohms.}$$

Since $R_2 = 0.25$ megohms

$\therefore R_1$ is 1 megohm.

6.17 Questions and problems.

1. Why is transmission through space not carried out at audio frequencies?
2. How is audio-frequency intelligence transmitted through space?
3. Define modulation.
4. What is meant by the modulation envelope?
5. Define percentage modulation.
6. What is meant by the modulation factor m?
7. Between what limits does percentage modulation vary?
8. Between what limits does the modulation factor m vary?
9. What is the disadvantage of modulation which is greater than 100%?
10. How can distortion result when there is less than 100% modulation?
11. Name the three principal methods of amplitude modulation.
12. Give the two main properties of amplifier and oscillator circuits which are necessary for amplitude modulation.
13. Draw a typical circuit for anode modulation.
14. Describe the operation of the circuit for anode modulation.
15. Explain why there must be a linear relation between the anode current and the anode voltage varying at the audio rate in anode modulation.
16. Why must there be a linear relation between the anode tank current and the anode voltage varying at the audio rate?
17. Why is anode modulation also called high-level modulation?
18. Between what limits is the anode voltage varied to obtain 100% modulation?
19. What is the source of the power required to generate the carrier wave?
20. Draw a typical circuit for grid modulation.
21. Explain the operation of the circuit for grid modulation.
22. Why is it not desirable to drive the grid positive during grid modulation?
23. What appearance should the dynamic $e_c - i_b$ curve have for minimum distortion?
24. Why is grid modulation also called low-level modulation?

25. Draw a circuit for cathode modulation.

26. Compare anode and grid modulations to show advantages and disadvantages of each.

27. Draw a typical circuit for suppressor-grid modulation.

28. Explain the operation of the circuit for suppressor-grid modulation.

29. How is the correct value of operating bias found for the circuit for suppressor-grid modulation?

30. When an r-f carrier is modulated by a sinusoidal audio frequency, how many frequencies are present in the resulting wave?

31. Are the resulting frequencies in question 30 a-f or r-f?

32. What are the resulting frequencies in question 30 called?

33. If a carrier of 600 Kc/s is modulated by an a-f wave of 400 c/s what are the frequencies present in the resulting wave?

34. How does the total power of the carrier and sidebands compare with the carrier power when there is 100% sinusoidal modulation?

35. Why is it desirable to keep the sideband power as high as possible?

36. What is the channel width in the broadcast band?

37. What supplies the power for the sidebands?

38. If a 10,000-watt carrier is present what sideband power is required for 100% modulation?

39. Name three types of transmission using amplitude modulation and define each one.

40. What is meant by back-wave in keying?

41. Why are the shapes of the r-f pulses forming the dots and dashes important?

42. Give two methods for suppressing higher harmonics in code pulses.

43. Draw a circuit for spark suppression at key contacts.

44. Give three factors which determine the position selected for a key in a transmitter.

45. What are the two general methods of keying?

46. Draw a circuit to illustrate primary power-supply keying and explain the operation.

47. Give an advantage and disadvantage of power-supply keying.

48. How is secondary power-supply keying carried out?

49. Draw a circuit to illustrate keying employing grid-controlled rectifier valves and explain the operation of the circuit.

50. Draw a circuit to show anode keying and explain the circuit operation.

51. How is cathode keying carried out?

52. Draw a circuit to illustrate grid keying and explain its operation.

53. Give an advantage of grid keying.

54. Draw a circuit to illustrate absorption grid keying and explain its operation.

55. Illustrate sinusoidally modulated r-f waves with
    (a) 100% modulation
    (b) less than 100% modulation
    (c) more than 100% modulation.

56. Give the formula used to make direct measurement of percentage modulation for the waveforms in question 55.

57. Illustrate trapezoidal patterns seen on an oscilloscope screen with
    - (a) 100% modulation
    - (b) less than 100% modulation
    - (c) more than 100% modulation.

58. An anode-modulated oscillator has a 100-watt valve. The valve works at an efficiency of 60% and is fully loaded when the aerial current is 80% modulated. The aerial power with this degree of modulation is 32% greater than the aerial power when the aerial current is unmodulated. The modulator valve gives a distortionless output and when its output is a maximum its efficiency is 20%. If the maximum power output of the modulator valve produces 80% modulation of aerial current find
    - (a) the maximum power output of the modulator valve
    - (b) the power dissipated as heat at the modulator valve anode when the microphone is quiescent
    - (c) the modulator valve rating
    - (d) the overall efficiency of the transmitter when its aerial current is unmodulated and when it is 80% modulated.

CHAPTER VII

DETECTION

7.1 Principles of detection (demodulation). As seen in the chapter on Modulation, it is not possible to carry out wireless transmission at audio frequencies directly due to several difficulties, some of which are

(i) the necessary aerials would be too long

(ii) aerials do not radiate efficiently at such low frequencies

(iii) only one station could be on the air at one time due to the resulting interference.

To overcome these difficulties the method of impressing the audio frequencies on a carrier wave of high radio frequency was devised. This modulated carrier-wave method of transmission has none of the above disadvantages, but in order for it to be useful it must be possible to reproduce the modulating audio frequencies at the point of reception. This process of recovering the modulating frequencies from the modulated r-f carrier wave is known as *detection* or *demodulation*.

In this chapter, only the detection of amplitude-modulated waves is discussed. The detection of frequency-modulated waves is discussed in Chapter XIII. In an amplitude-modulated carrier wave the frequencies present are the upper sidebands, the carrier frequency and the lower sidebands. These are all r-f frequencies. If the corresponding r-f voltages are applied directly to a loudspeaker or earphones no sound is heard. The alternate half cycles, more or less equal, succeed one another too rapidly for the diaphragm to follow. The diaphragm responds only to the average value which is zero since the voltage waveform is symmetrical. If, however, the waveform is distorted in some way, frequency components other than the carrier and sidebands are obtained and one of these components will be at the frequency with which the carrier was modulated at the transmitter. If the r-f voltage, having this distorted waveform, is applied to a loudspeaker or earphones the diaphragm responds to the average value which varies according to the modulation envelope. Any asymmetrically-conducting device which provides complete or partial rectification is suitable for distorting the wave and is called a *detector*. That is, a detector produces an audio-frequency output voltage which has the same waveform as the modulation envelope.

7.2 Diode detection. The diode detector is the simplest and most widely used of the various types of detectors. It derives its name from the fact that a diode valve is used as the rectifier which distorts the incoming r-f signal. The basic circuit is given in Fig. 7-1.

A. Operation of the Circuit

The desired r-f signal is selected by means of the tuned circuit and applied

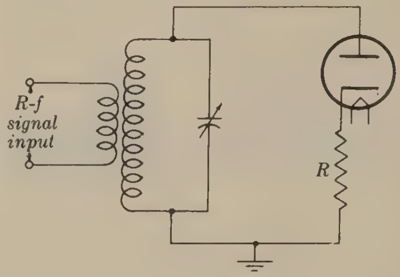


FIG.7-1 Simple diode detector circuit without a filter.

across the diode and the resistor R in series. The waveform of this applied voltage is shown in Fig. 7-2 (a). The section from A to B shows an r-f carrier voltage and from B to C a carrier which is modulated at a single audio frequency.

When the r-f carrier voltage is applied across the diode and the resistor, the diode conducts on the half cycles when the anode is positive. Therefore pulses of current flow up through the resistor R producing a voltage across R as shown between A and B in Fig. 7-2 (b). When the modulated r-f voltage is applied across the diode and resistor, the amplitudes of the current pulses through R vary with the changing amplitudes of the positive half cycles of input voltage. The resulting voltage across R is shown between B and C in Fig. 7-2 (b). If the voltage across R, shown between A and B in Fig. 7-2 (b), is applied to a loudspeaker or earphones no sound is heard since the average value of voltage is steady. However, when the voltage between B and C is applied a sound is produced since the average value of voltage varies in accordance with the modulation envelope. Thus the sound which was used to produce the modulating signal is heard.

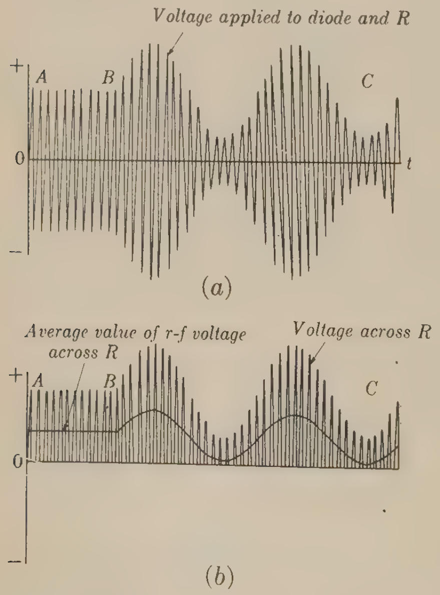


Fig.7-2 Voltage waveforms in diode detector (a) R-f input to the detector (b) Output voltage of the detector.

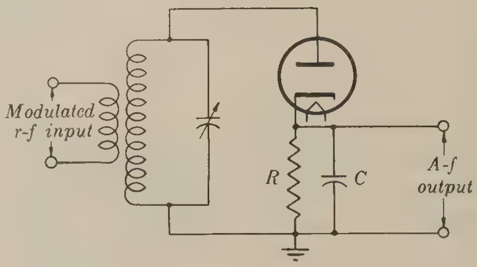


Fig.7-3 Diode detector circuit with a filter.

The a-f variations form the useful voltage output from the circuit. Thus, a capacitor C is placed in parallel with the resistor R as in Fig. 7-3, to eliminate the r-f variations which are of no further use.

When the r-f carrier is applied, the diode conducts on a positive half cycle and C charges up to practically the peak value of the applied voltage. As the anode voltage falls and goes negative for a half cycle the diode does not conduct. The charge on C then begins to leak off through R, but the size of R is chosen so that only a small amount of charge leaks off by the time that the voltage again becomes positive enough to make the diode conduct and recharge the capacitor to the peak voltage. The resulting voltage across C and R is therefore a steady d-c at practically the peak voltage of the incoming signal with a slight r-f ripple superimposed on it, as shown in the section AB of Fig. 7-4 (a). The voltage pulses under the capacitor voltage are those obtained when C is not present. Now if the r-f is modulated, its peak voltage varies at an a-f rate. However, as before, the capacitor continues to charge to the peak voltage and discharge slightly through R when the anode is negative with respect to the cathode. The resulting voltage across

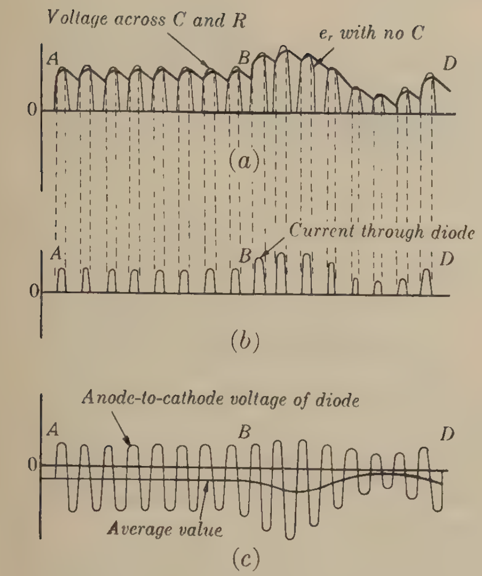


FIG.7-4 Voltage and current waveforms in a diode detector with filter (a) Voltage developed across RC filter (b) Current through the diode (c) Voltage across the diode.

C and *R* is an a-f voltage with a slight r-f ripple superimposed on it as shown in section *BD* of Fig. 7-4 (a). The values of *R* and *C* are chosen so that although the voltage across *R* is practically the peak voltage of the r-f, yet the *RC* time constant is such that the voltage across *R* can follow the a-f variations of the r-f voltage amplitude. The magnitude of the r-f ripple is greatly exaggerated in this diagram, since, in actual practice, there are hundreds of r-f cycles during one a-f cycle. Thus the voltage across *R* varies in accordance with the a-f modulating signal and detection has taken place. Since the capacitor *C* charges to the peak value of the r-f pulses, the amplitude of the a-f output is greater than that obtained when no capacitor is used. In the latter case the a-f output is only the average value of the r-f pulses.

The anode-to-cathode voltage across the diode is shown in Fig. 7-4 (c). On the positive half cycles the diode conducts and its impedance is low. Thus a small amount of the applied voltage appears across the diode. On the negative half cycles the diode does not conduct and its impedance is very high. Therefore, all of the applied voltage appears across the diode. The r-f waveform of voltage across the diode is distorted so that the average value varies according to the modulation envelope.

Current through the diode during the same period *A* to *B* to *D* is shown in Fig. 7-4 (b). The current pulses have a small amplitude since the diode conducts only when the input voltage on the positive half swing exceeds the voltage across the capacitor.

In Fig. 7-4 (a) it is seen that the voltage across *R* is a varying d-c. That is, it may be considered to be made up of two parts, a steady d-c and an a-c component. It is not desirable to pass the d-c voltage on to the next stage and therefore a blocking capacitor C_1 (Fig. 7-5) is used when coupling the output a-f voltage from the detector to the grid of the following amplifier stage.

The operation of the circuit is also explained by using the fact that voltages which are not sinusoidal are made up of sinusoidal components of various frequencies. The incoming r-f voltage appears almost entirely across the diode since *C* is chosen so that X_c is much less than *R* at r-f. The incoming r-f is rectified by the diode. The resulting distorted diode current is made up of r-f and a-f

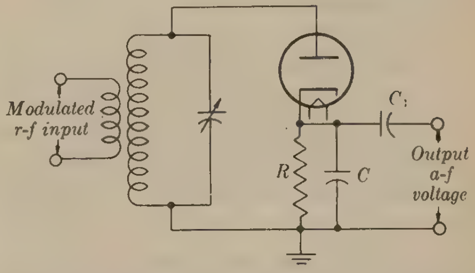


FIG.7-5 Diode detector and output coupling circuit.

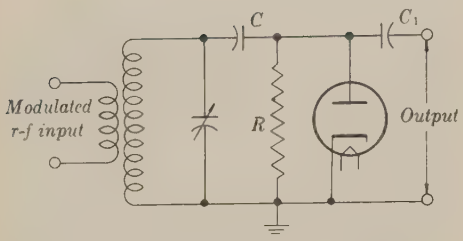


FIG. 7-6 Diode detector with parallel load.

components and a d-c component. R is needed to provide a path for the d-c. The value of C is such that it by-passes the r-f but offers a large reactance to the a-f and therefore the modulating signal voltage appears across it. R is equal to or greater than X_C at a-f so that it does not shunt the capacitor and therefore lower the a-c load. Finally, the a-f output voltage is coupled to the next stage through a blocking capacitor C_1 which has a low reactance to a-f.

B. Other Forms of the Circuit

Fig. 7-5 shows the most widely used form of the basic diode detector circuit. However, the basic circuit may appear in some slightly different arrangements which are considered here.

Fig. 7-6 shows a parallel arrangement of the diode and RC filter circuit. The operation is similar to the previous case in which the diode and RC circuit are in series. However, the output is the voltage across the diode as shown in Fig. 7-4 (c). Since this output consists of r-f and a-f components, additional r-f filtering is required.

In the two previous circuits the diode provides half-wave rectification. By using the double diode, full-wave rectification is obtained. The circuit is shown in Fig. 7-7. The input and output voltage waveforms are shown in Fig. 7-8. Since each diode conducts on alternate half cycles of the input voltage, pulses at double the r-f frequency appear across R in the absence of C, as shown in Fig. 7-8 (b). The advantage of this circuit is that the r-f ripple in the a-f output is considerably reduced. The r-f ripple is double the frequency of the incoming r-f and therefore the reactance of C at this frequency is very small. As a result there is no need for any further r-f filtering in the a-f output. This advantage is offset to some extent by the additional expense and also by the fact that since the transformer is centre-tapped, only half the incoming voltage is applied to each diode. This reduces the sensitivity of the detector.

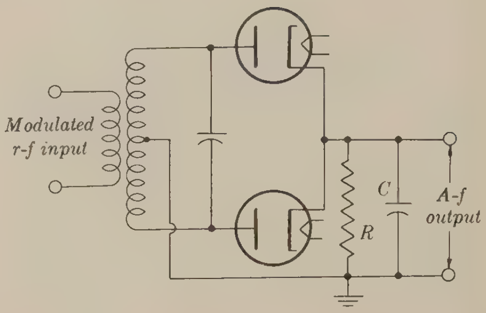


FIG. 7-7 Full-wave diode detector.

C. Characteristics of the Detector

1. LINEARITY.

By *linearity* or *fidelity* is meant the ability of the detector to produce as an a-f output, a voltage which is directly proportional to the modulation envelope of the applied signal. The linearity of a diode detector is very good. This is due mainly to the fact that the dynamic $e_b - i_b$ curve (Fig. 7-9) is very straight over most of its length. To reduce any distortion caused by the curved portion at the bottom, the input voltage should be as large as possible, usually not less than $10V$ peak. In addition the percentage modulation should be kept slightly under 100% since with 100% modulation the modulation envelope dips down to cause operation in the curved portion of the $e_b - i_b$ curve.

A perfectly linear valve curve is not

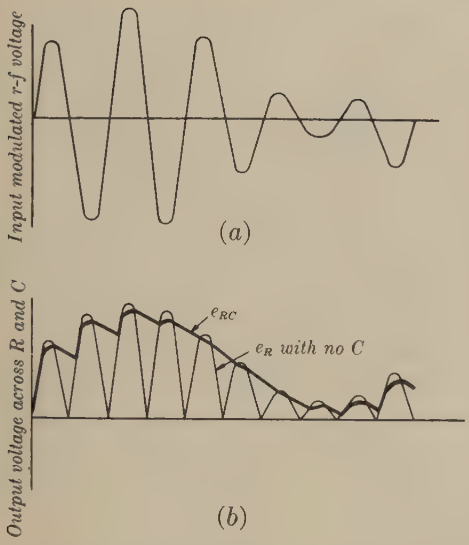


FIG.7-8 Voltage waveforms in a full-wave detector (a) Modulated r-f input (b) Output voltage.

enough in itself to guarantee no distortion. The values of the load resistance R and the capacitor C are an important consideration. If C is too large, then the time constant CR is too long compared to the time of an a-f cycle. The capacitor C is not able to discharge quickly enough as the a-f voltage decreases from its maximum amplitude. This type of distortion is called *diagonal clipping* (Fig. 7-10). As far as the value of R is concerned, two factors must be kept in mind. Firstly, the effective a-c load to the a-f and not just the value of R is the factor which determines the amount of distortion. That is, the grid resistor of the following stage must be sufficiently large to prevent a-f shunting of R. It is usually made at least twice the value of R. Secondly, R must be much greater than the r_a of the diode.

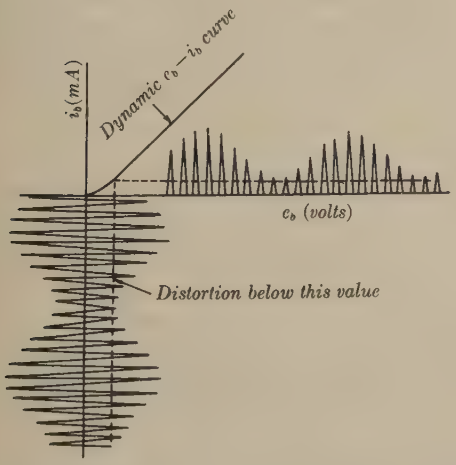


FIG.7-9 Diode $e_b - i_b$ curve showing linear operation.

2. INPUT IMPEDANCE OR RESISTANCE.

A diode detector is characterized by a low input impedance since it draws current from the source. This is a disadvantage since it lowers the Q of the tuned input circuit, thus reducing its selectivity. This is particularly troublesome if there are not several stages having good selectivity preceding the detector.

3. SENSITIVITY.

Another characteristic of the diode detector is its poor sensitivity. It cannot detect signals of less than approximately one-quarter of a volt. However in most modern receivers this is not a problem since one stage of r-f amplification can give an output of $10\,V$ peak. As a result the diode detector is widely used.

The poor sensitivity of the diode detector results from two facts. Firstly, the a-f output voltage is smaller than the input modulation amplitude. The ratio

$$\frac{\text{a-f output amplitude}}{\text{modulation amplitude}} \times 100$$

is called the *detector efficiency* and is between 90-95% for a well designed diode

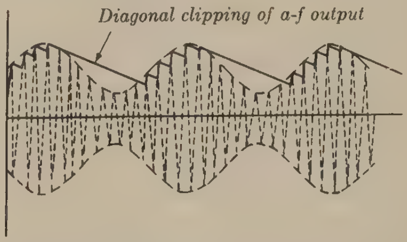


FIG.7-10 Diagonal clipping due to excessively large filter capacitor.

detector. This lack of amplification is the main reason for the poor sensitivity. Secondly, however, the low input impedance to the diode reflects a low impedance into the primary of the tuned input circuit and the r-f voltage developed here is lower than if the input impedance were high.

4. Signal-Handling Capabilities.

The diode detector is able to handle very large input signals without overloading and causing distortion due to the long linear portion of the $e_b - i_b$ curve. This is one of the advantages which has brought the diode detector into general use.

5. Frequency Response.

The frequency response of a diode detector is excellent. That is, for a given input voltage, the output voltage remains constant over a wide range of frequencies.

6. Automatic Volume Control.

The diode detector circuit is particularly useful in that it provides a method of obtaining automatic volume control in a radio receiver without using a separate circuit.

D. The Practical Circuit

The values of R and C used in practice depend on the frequency range of the receiver in which the detector is used. As a practical example let us consider the values required in a receiver to be used over the broadcast band, 500 to 1500 Kc/s. The value of R, in Fig. 7-3, must be much greater than the r_a of the diode and also approximately equal to the reactance of the capacitor C at a-f. Values ranging from 100,000Ω to 1$M\Omega$ are used, with 300,000Ω being a very common value. The value of C must be chosen so that the CR time constant will be long compared to the time of an r-f cycle and short compared to an a-f cycle. The highest audio frequency encountered in ordinary broadcast receivers is about 5000 c/s. A time constant frequently used is about $\frac{1}{4}$ of an a-f period, that is $\frac{1}{20{,}000}$ sec.

$$\text{Hence} \quad CR = \frac{1}{20{,}000} \text{ sec.}$$

$$\text{If} \quad R = 300{,}000\Omega$$

$$\text{Then} \quad C = \frac{1}{20{,}000 \times 300{,}000}$$

$$= \frac{1}{6 \times 10^{9}}$$

$$\doteq 150 \times 10^{-12}$$

Therefore C is approximately $150\mu\mu F$.

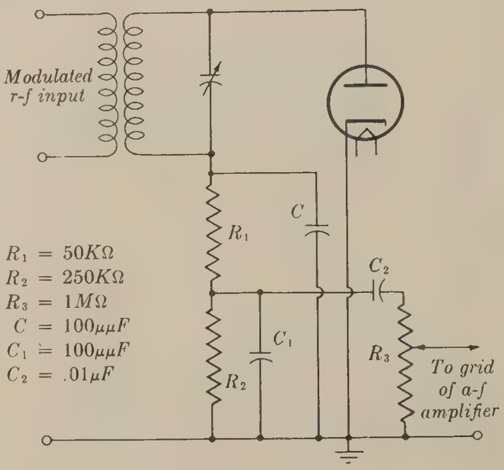


Fig.7-11 Practical diode detector.

A common arrangement of a practical diode detector circuit is shown in Fig. 7-11. The usual load resistor R is broken up into R_1 and R_2. In order to reduce any r-f voltage in the output to a minimum the capacitor C_1 is added in the position shown and the output is taken across R_2 instead of across both R_1 and R_2. C_2 is the blocking capacitor to stop the d-c voltage from affecting the grid of the next stage. It must have a small reactance to a-f and is therefore of the order of $.01\mu F$. R_3 is large compared to R_1 and R_2 to minimize any a-c shunting of the diode load. R_3 is a potentiometer to vary the amplitude of the a-f output and so act as the volume control in a receiver.

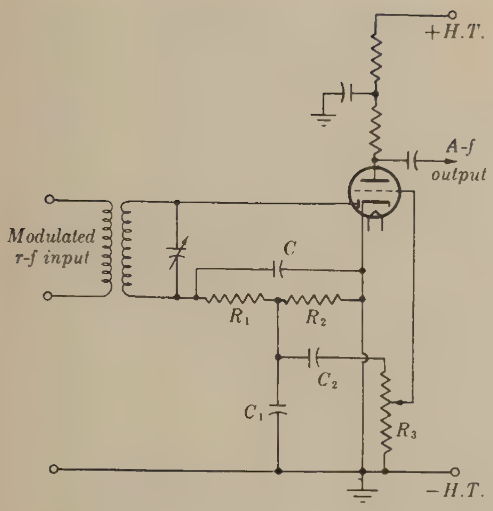


FIG.7-12 Single-valve diode detector and a-f amplifier.

It is very common practice to include the diode detector and the succeeding a-f triode amplifier in one valve. A circuit making use of such a valve is shown in Fig. 7-12. The numbering of the components corresponds to that in Fig. 7-11 so that comparison may be made.

7.3 Crystal detection. Another form of rectifying device which is used for detection consists of a crystal in contact with a metal, or combinations of two crystalline substances in contact with each other. Some crystals having the unique property of conducting better in one direction than the other are galena, silicon, carborundum, zincite, iron pyrites, and hertzite. The crystal is rigidly mounted in a metal base and either one of the sharp angles of the crystal presses against a metal plane, or a fine metal point bears against the crystal. With galena only a light pressure is needed and a light, pointed wire called a *cats-whisker* pressing on the crystal is commonly used. Silicon and iron pyrites require a slightly higher pressure which makes them more stable but at the same time less sensitive in rectifying small signals. Carborundum can stand a still higher pressure making it very stable but correspondingly less sensitive. Besides the fact that the sensitivity is very dependent on the pressure, it also varies widely with different spots on the crystals and the contact must be placed so that it is pressing on one of the sensitive spots. Two crystal-voltage crystal-current curves are shown in Fig. 7-13.

When a crystal is used as a detector the circuit is the same as for the diode detector except that the diode valve is replaced by the crystal. A circuit is shown in Fig. 7-14. The arrow in the crystal symbol points oppositely to the direction of best conduction of electrons. The action is identical to that of a diode detector. *C* is an r-f by-pass capacitor so that most of the r-f voltage appears

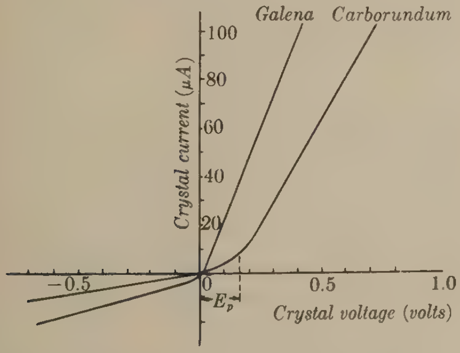


FIG.7-13 Voltage-current characteristics for two crystals.

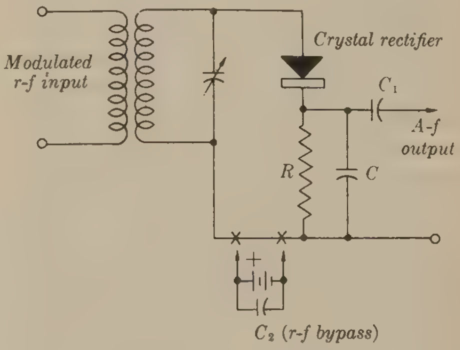


FIG.7-14 Crystal detector.

across the crystal. C and R are also chosen so that they have a long time constant compared to the length of an r-f cycle but a short time constant compared with an a-f cycle. For certain conditions it is desirable to adjust the operating point to a point of maximum curvature of the rectifier characteristic. This is done by means of a positive polarizing voltage E_p as shown in Fig. 7-13. In such a case a small battery in series with the crystal and by-passed for r-f is inserted in the circuit at the points marked X as shown in Fig. 7-14.

The actual resistances of the crystals vary widely with size, type and length of time used, but the important factor to be considered is the rectifying ratio which is the resistance in one direction divided by the resistance in the opposite direction. An average value is 10 to 1 but in a very good crystal it may be as much as 40 to 1, for example, 12,000Ω in one direction and 300Ω in the other. A crystal is of no more use if the ratio falls below 4 to 1 and it is then said to be *burnt out*. A sensitive point on a crystal can be very easily burnt out by putting too large a signal on the crystal and causing too much d-c to flow. Crystals also have the disadvantages of being quite delicate mechanically and variable in their properties. They have, nevertheless, several advantages over valve rectifiers. The circuit is simpler, there are no filaments to heat and no replacements necessary due to filaments burning out. They are very compact and have a sensitivity which compares quite favourably with diode detectors. Their main advantage lies in the fact that there are no transit time effects as in a valve and they have an extremely low capacitance between the terminals. This makes them very useful, if not indispensable, in very-high and ultra-high-frequency work. This application is discussed in Chapter XVII.

7.4 Linear detection and square-law detection. An ideal detector is designed to produce an output exactly proportional to the size of the modulating signal. Unfortunately, many detectors fail to do this and distortion results. They often introduce currents of frequencies which were in neither the received signal nor the modulating intelligence and thus give rise to amplitude distortion. If the detector does not respond equally well to all the modulating frequencies frequency distortion is introduced. And lastly, if the detector reproduces the different frequency components in altered phase relationships to the original, then phase distortion is introduced. It should be emphasized again that distortion of the r-f carrier wave is absolutely necessary, but distortion of the a-f modulating signal in any of the previously mentioned ways is undesirable. Although no detector in practice is free from such distortion, in all practical detectors the degree of distortion present is quite tolerable.

In classifying methods of detection it is not sufficient to name them only according to the type of rectifier used in the circuit. Since the strength of the input signal and the type of relationship between the input and output voltages are inherently involved in the detector operation, these characteristics provide another method of classification. As a result detectors are also described as *linear* or *square-law detectors.* Linear detection is always associated with large input signals and consequently is sometimes called *power detection.* On the other hand, square-law detection is only used with small input signals and may therefore be called *weak-signal detection.* As its name indicates, linear detection makes use of the maximum straight portion of a voltage-current curve so that the a-f output is as nearly directly proportional to the modulation envelope of the input as possible. That is, there is a linear relationship between input and output. Linear detection, therefore, is the more desirable type if lack of distortion in the a-f is an essential feature. With linear detection, the distortion of the symmetrical r-f input necessary to produce detection, is obtained by complete rectification. The types of detectors discussed thus far in the chapter are examples of linear or power detectors. Square-law detection makes use of the curved portion of the current-voltage curve to produce the distortion necessary for detection. As a result the output is proportional to the square of the input. The reason for this is that the curved region at the bottom of any valve current-voltage curve is closely represented by a second degree equation of the form $y = ax^2+bx+c$.

A simple example of the way in which square-law detection takes place is demonstrated by assuming that the curved section of a current-voltage curve is given by $4y = x^2$, where y is the current through the rectifier and x is the applied voltage. In this equation $y = \frac{1}{4}x^2$. Hence y is directly proportional to x^2. The graph of this equation is given in Fig. 7-15. Square-law detectors always operate at some point in the middle of the curved region, at the point O in this diagram. From the graph it is seen that if the input voltage swings to the right of the point O, the change in current is larger than if it swings to the left of the point O. This means that a symmetrical r-f input voltage results in a distorted rectifier current and detection takes place. However, this method also introduces undesirable distortion of the a-f modulating signal, especially if the input has a high percentage of modulation since the average value of current does not vary exactly with the modulation envelope. The percentage distortion of the a-f is given by the formula

$$\frac{m}{4}\times 100$$

where m is the degree of modulation.

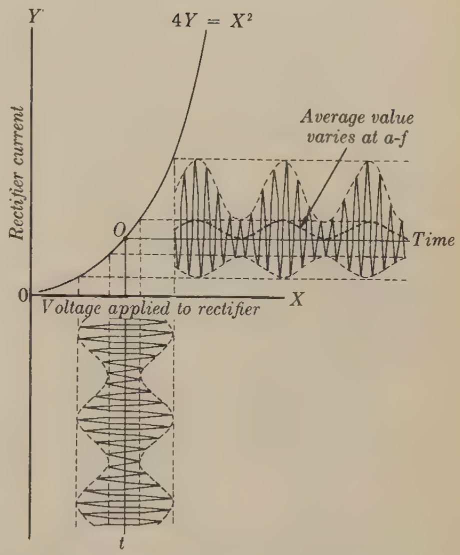


FIG.7-15 Voltage-current relationships in a square-law detector.

The maximum percentage distortion is obtained with 100% modulation and is equal to 25%. Square-law second detectors were commonly used when only weak signals were available for the detector because of a lack of good r-f amplifiers and when the percentage modulation was not very high. However, square-law detection is still of importance for several reasons.

(i) Due to the curvature at the bottom of all $e_c - i_b$ and $e_b - i_b$ curves, any detector operates on the square-law principle if the applied signal becomes sufficiently small.

(ii) It gives a distortionless output of the difference frequency for heterodyne signals, which is of importance in superheterodyne receivers.

(iii) It is useful in measuring instruments such as vacuum-tube voltmeters because their behaviour can be simply and exactly predicted mathematically when the square-law detector is used.

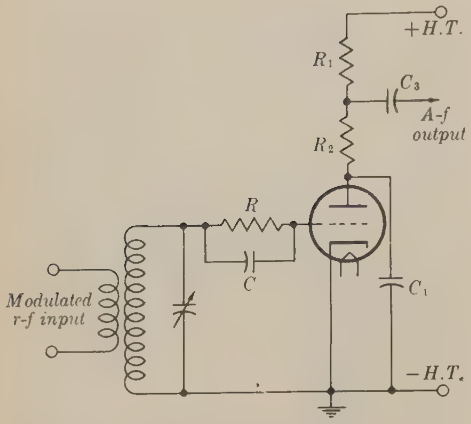


FIG.7-16 Cumulative-grid detector with series grid-leak bias.

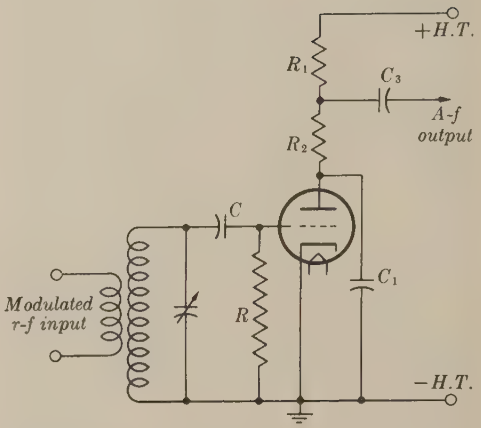


FIG.7-17 Cumulative-grid detector with parallel grid-leak bias.

Anode-bend detectors, cumulative-grid detectors or crystal detectors are all sometimes used as square-law detecting devices. In the case of the latter it may be necessary to adjust the operating point to the optimum position on the characteristic curve by means of a small polarizing potential as shown in Figs. 7-13 and 7-14.

The principles of detection may be summed up by saying that the requirements for a detector are as follows.

(i) Any detector must be a rectifier, partial or complete.
(ii) Any detector circuit must include an impedance across which is developed a voltage at the modulating frequency.

When such requirements are met, the results are

(i) the production of a non-symmetrical output from a symmetrical input variation
(ii) the selection and rejection of a-c components in the asymmetrical output by the filter circuits
(iii) a direct-current component. This is made use of in automatic volume control circuits which are discussed in Chapter IX.

These fundamentals apply equally to *first* or *second detectors*. In this chapter only the various types of second detectors are considered. For a discussion of first detectors see Chapter IX on superheterodyne receivers.

7.5 Cumulative-grid detection. This method of detection is also called *grid-leak detection* or *grid-current detection.* It employs a triode valve and provides amplification as well as detection. For this reason it was widely used when good r-f amplification did not exist. With this method, the detection takes place in the grid circuit. Figs. 7-16 and 7-17 show the two possible arrangements in the grid circuits but the operation is the same for both. Audio transformer coupling may also be used in the output since it increases the gain of the detector stage.

To understand the operation of a grid-leak detector it must be considered in two parts, one, the grid circuit in which the detection takes place, and two, the anode circuit of the triode which provides amplification and additional filtering. The action in the grid circuit corresponds with that in the diode detector. The grid is the diode anode and *R* and *C* make up the diode load. The choice of *R* and *C* depends on the same considerations as were discussed under diode detection. The incoming r-f voltage is applied between grid and cathode which form the anode and cathode of the diode and is rectified. The a-f component in the distorted r-f voltage appears across the RC circuit. These a-f voltage variations are applied between grid and cathode and cause the anode current to vary accordingly. An amplified a-f output voltage is developed across the resistors R_1 and R_2. The r-f voltages which were by-passed to ground by C are also amplified by the triode. These are by-passed to ground by C_1 which offers a much lower reactance to r-f than does the anode load R_1 and R_2. Thus only the a-f voltage is passed on to the next stage to be amplified. This general description of the operation holds for both linear and square-law grid-leak detection.

Grid-leak detection is used with both weak and strong signals. Although the circuit and general operation is the same in each case, the detailed operations differ sufficiently to be discussed under separate headings.

A. Weak-Signal Grid-leak Detection

The rectification in the grid circuit is caused by the curvature in the dynamic grid-voltage grid-current $e_c - i_c$ curve (Fig. 7-18). This curve indicates that grid current does not cut off until the grid is one volt negative with respect to the cathode. A bias is developed across the grid leak resistor due to grid current flow. With weak signals of the order of 0.1 to 1.0 volts this bias is such as to make the operating point come in the middle of the curved region. The r-f variations are not large enough to drive the grid beyond cut-off and therefore grid current flows at all times.

With zero signal, grid current flows through R and charges C (Fig. 7-16) and develops a bias across C and R which sets the operating point at A in Fig. 7-18. When the modulated r-f voltage is applied to the input of the circuit, the grid current rises on the positive half cycles of the r-f input and C is charged to nearly the peak of the input voltage.

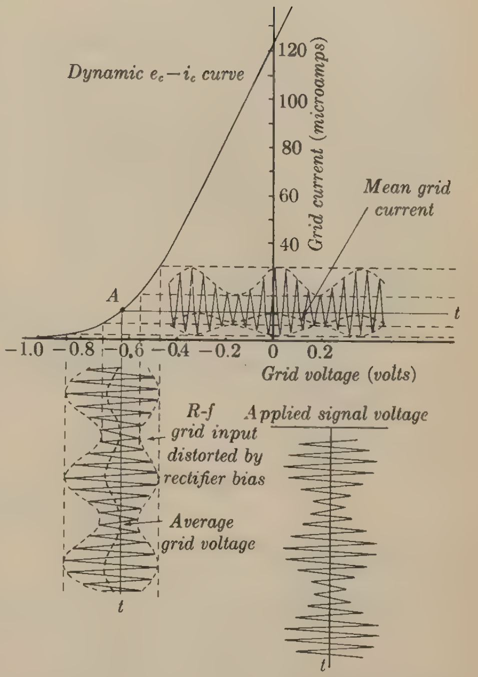


FIG. 7-18 $e_c - i_c$ curve illustrating weak-signal grid-leak detection.

On the negative half cycles the grid draws very little current and C begins to discharge through R. Since R and C are chosen to have a long time constant compared to the time of an r-f cycle, the voltage across R remains practically the peak voltage of the r-f input. However, the time constant is short compared to the time of a cycle of the modulation envelope. Hence the bias across C and R varies with the modulation envelope. Since the r-f input varies about this bias changing at an a-f rate, the voltage between grid and cathode is distorted as shown in Fig. 7-18. Therefore, the anode current of the valve is distorted in a similar manner as in Fig. 7-19. The average value of anode current provides the a-f output.

Since this detector operates entirely over the curved portion of the $e_c - i_c$ curve, a square-law relationship exists

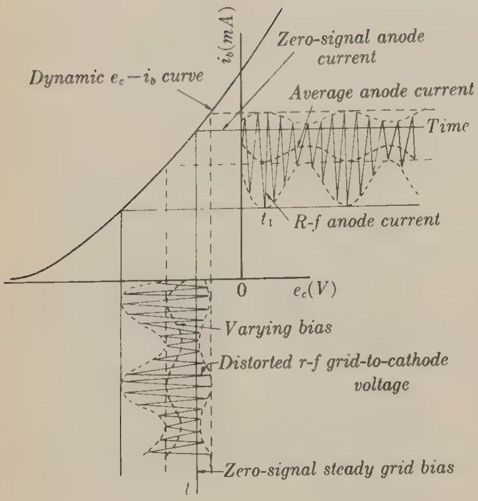


FIG. 7-19 Anode-current, grid-voltage waveforms for grid-leak detection.

between the modulation envelope of the input and the a-f output voltage.

The distortion in the a-f is quite severe and depends on the percentage modulation in the input. To prevent any further a-f distortion in the anode circuit when amplification takes place, it is necessary that the triode be operated over a linear portion of the $e_c - i_b$ curve (Fig. 7-19). The length of the linear portion is controlled by the value of H.T. voltage. Since the signal voltage and grid current are very small, large values of R and C are required to produce satisfactory bias and sensitivity. Typical values of C and R are $300 \mu\mu F$ and $1-3 M\Omega$, while the d-c anode voltage on the triode is about $50 V$.

The linearity of this detector is poor owing to the square-law operation. Another possible source of frequency distortion is the r-f by-pass capacitor C_1 which, if it is too large, will also by-pass the higher audio frequencies. On account of the poor fidelity, this detector cannot be used for r-t work, but it is quite satisfactory for c-w or m-c-w, where distortion of the code signal is of less consequence. A second disadvantage of this detector lies in the fact that it draws grid current, which results in a damping of the tuned input circuit with a consequent loss of selectivity and sensitivity in the preceding stage. Besides this it can handle only a small range of signal strengths without overloading and causing very severe distortion. Although widely used at one time, it has a limited application today.

B. Power Grid-Leak Detection

As its name indicates, this method of detection is very similar to the preceding one except that much larger signals up to $5 V$ or more are applied. With the application of larger input signals the operating point P on the $e_c - i_c$ curve moves back almost to cut-off (Fig. 7-20). On the negative half cycle the grid is driven well beyond cut-off and no current flows from cathode to grid. On the positive half cycles, pulses of grid current flow to keep C charged to the peak of the r-f input voltage. Again, this flow of grid current has the disadvantage of reducing

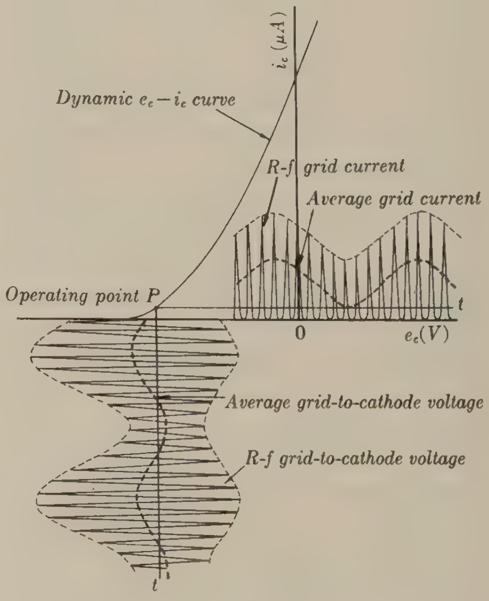


FIG. 7-20 $e_c - i_c$ curve illustrating power grid-leak detection.

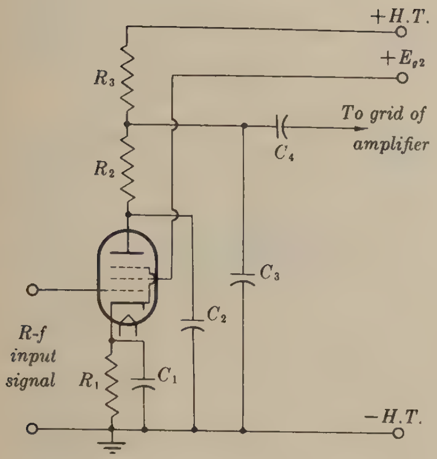


FIG.7-21 Anode-bend detector.

the selectivity of the input circuit. Since the input voltage is so much larger, the positive half cycles operate mostly over the linear portion of the $e_c - i_c$ curve and the a-f component across R approaches fairly closely in waveform the input modulation envelope. Because the a-f variations on the grid of the triode are also much larger, it is necessary to increase the length of the linear portion of the triode $e_c - i_b$ curve if distortion is to be avoided. This is accomplished by increasing the triode d-c voltage from $50V$ to about $150V$.

Since the signal amplitude is large, smaller values of R and C than were required for weak-signal detection provide the desired bias and sensitivity. Typical values of R and C are $\frac{1}{4}M\Omega$ and $100\mu\mu F$. Although the linearity of this detector is much better than for the weak-signal type and larger signals are applied to this detector it still falls far short of the diode detector in these respects. The main advantage of this circuit lies in its sensitivity. It finds application in regenerative receivers.

7.6 Anode-bend detection. This method of detection is also called *plate detection* or *grid-bias detection.* It makes use of the curvature of the $e_c - i_b$ curve of a triode or pentode to produce rectification. The detection, therefore, takes place in the anode circuit and a certain amount of amplification is obtained along with the detection. It is possible to use the curvature which occurs either near cut-off or near saturation to provide the rectification. If the former is used then the method is called *lower-anode-bend detection* while if the latter is used it is described as *upper-anode-bend detection.* The circuit is the same for either method but the operating conditions are changed somewhat. A circuit is given in Fig. 7-21.

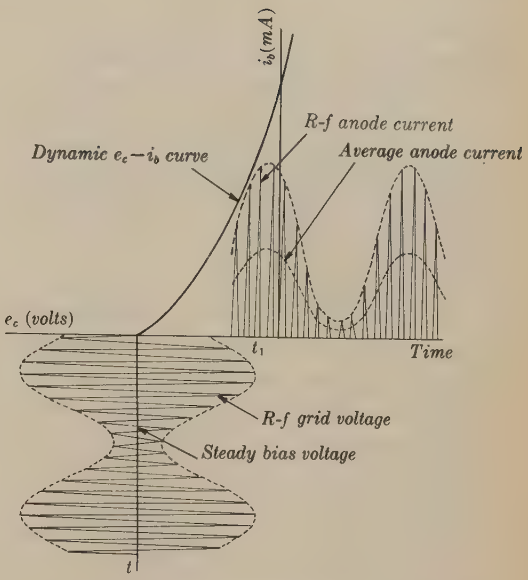


FIG.7-22 $e_c - i_b$ curve illustrating lower-anode-bend detection.

The operation of a linear anode-bend detector using lower-bend detection is understood from a study of Fig. 7-22 which shows an $e_c - i_b$ curve with the input grid voltage and output anode current waveforms. The valve is biased practically to cut-off point. In the circuit (Fig. 7-21) cathode bias is used. Since C_1 is a by-pass capacitor for both r-f and a-f, a $25\mu F$ electrolytic in parallel with a $500\mu\mu F$ mica capacitor is commonly used. The r-f input from the preceding stage is applied directly between grid and cathode. The grid potential then varies symmetrically on either side of the fixed bias. During the positive half cycles of the input voltage, current flows through the valve; during the negative half cycles, the input voltage swings the grid below cut-off so that no current flows.

The anode current flows in pulses and is therefore unsymmetrical and has an average value that varies in accordance with the a-f modulation. The anode load also forms a filter circuit which by-passes the r-f to ground. An amplified a-f voltage is developed across R_3 and is coupled to the next stage by C_4. The r-f filter capacitors increase the amplitude of the a-f output by charging to the peaks of the r-f pulses. The choice of R_2C_2 and R_3C_3 depends on the same criterions as were applied in the case of the filter circuit in the diode detector. C_4 is the coupling capacitor which offers a low reactance to a-f.

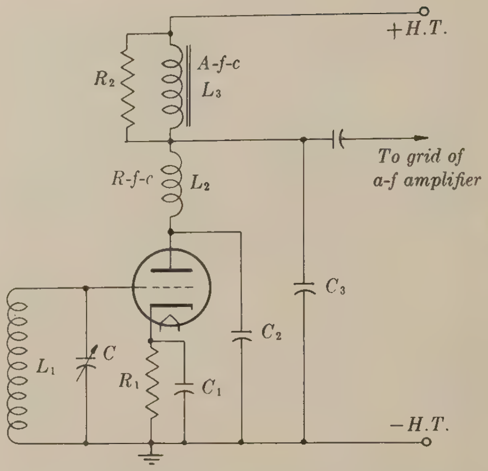


FIG.7-23 Anode-bend detector with inductive loading.

Fig. 7-23 shows a different load and filter combination in the anode circuit. The amplified a-f voltage is developed across L_3R_2. The a-f choke L_3 is shunted by R_2 to improve the a-f response at low frequencies. L_2 is an r-f filter choke. With large signals the linearity is fairly good but the anode load must be large to minimize the effect of the curvature at the bottom of the $e_c - i_b$ curve. The choice of the r-f by-pass capacitors C_2 and C_3 is critical. If they are too large the higher audio frequencies are by-passed but if they are too small the amplitude of r-f in the anode circuit is too large and the negative feedback through the anode-grid capacitance becomes appreciable and damping of the input circuit occurs. Another consideration in the prevention of distortion is that the input signals must never be large enough to drive the grid positive. If this occurs the grid draws current and two undesirable effects result. One is the reduction of selectivity of the input circuit. The other is a distortion of the a-f output. This second effect is caused by the grid rectification produced when the grid current flows. From a comparison of the anode currents in Figs. 7-19 and 7-22 it may be seen that at time t_1 on the graph there is a decrease in the average anode current with grid rectification while there is an increase in average anode current with plate rectification. Therefore if the grid is driven positive during the modulation peaks the effect of grid rectification counteracts the effect of anode rectification and the positive pulses of anode current are clipped during the modulation peaks. Thus the negative peaks of a-f output voltage are clipped. Since under normal operation the grid is not driven positive and no grid current flows there is a very high input impedance with a resulting high selectivity in the input circuit. The input impedance is not infinite due to the shunting effect of grid-to-cathode capacitance which prevents the use of this circuit above about 30 Mc/s. This detector has the advantage over the diode of higher sensitivity and selectivity but has only fair linearity and signal-

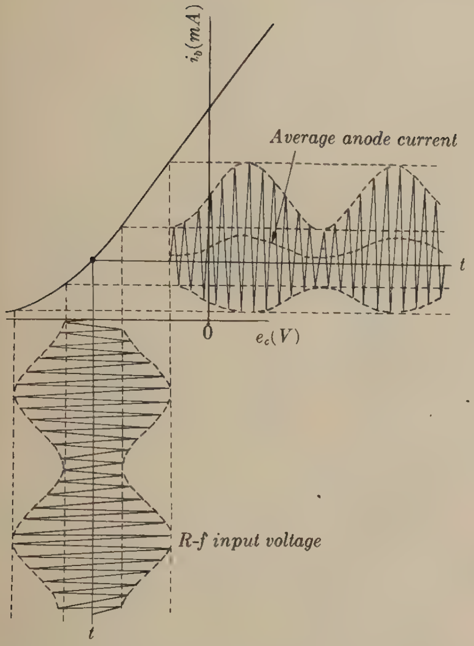


FIG.7-24 $e_c - i_b$ curve showing lower-anode-bend square-law detection.

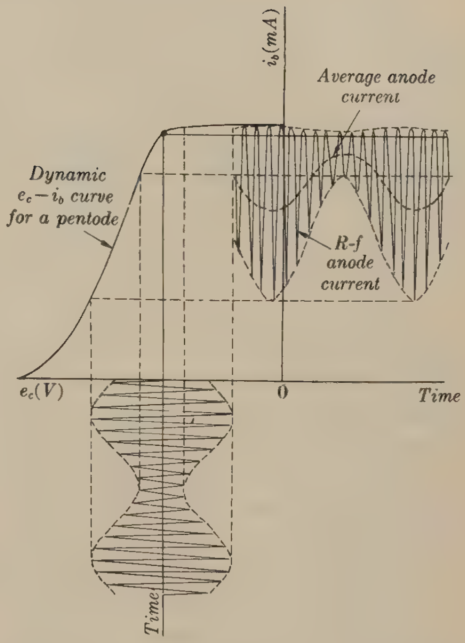


FIG.7-25 $e_c - i_b$ curve illustrating upper-anode-bend detection.

handling capabilities. It is used in inexpensive broadcast receivers and in amateur communication sets where fidelity is not a primary consideration.

The lower-anode-bend detector is sometimes used as a square-law detector. The circuit is exactly the same as above except that only small signals are applied and the grid bias is adjusted to set the operating point in the middle of the curved region of the $e_c - i_b$ curve as in Fig. 7-24.

Upper-bend detection is only used with pentodes because the upper bend occurs in the region of negative grid potentials while with triodes the upper bend occurs in the region of positive grid potentials. The circuit is the same as in Fig. 7-21 except the grid bias is adjusted so that the anode current is at saturation. The graph showing how the rectification takes place is given in Fig. 7-25. Upper-bend detection with pentodes gives slightly stronger signals but has the disadvantage of consuming more power from the supply than lower-bend detection.

7.7 Infinite-impedance detection. This detector operates as a lower-anode-bend detector, except that the anode load R_1 is in the cathode circuit as in Fig. 7-26. R_1 is chosen so that when no signal is being applied the d-c anode current biases the valve very close to cut-off. R_1 is by-passed for r-f, but not for a-f, by C_1. C_3 and R_3 maintain the anode at both r-f and a-f ground potential.

Since the valve is biased close to cut-off anode current flows in pulses during the positive half cycles of the modulated r-f input as in Fig. 7-27. The average value of the anode current varies in accordance with the modulation envelope. Thus, since C_1 is not an a-f by-pass, the voltage across C_1R_1 also varies in accordance with the modulation envelope. This a-f voltage is the output of the detector and also provides negative feedback in the form of an additional bias which varies at the a-f rate. The r-f

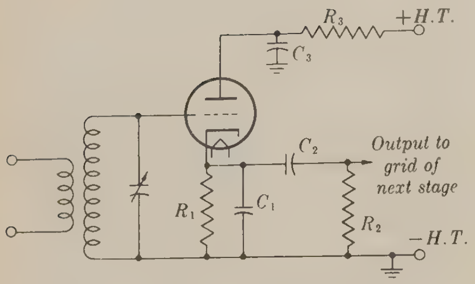


FIG.7-26 Infinite-impedance detector.

input varies about this a-f bias and appears as in Fig. 7-27. Since this bias increases as the signal amplitude increases and since the initial bias is large the grid never goes positive. Since the grid never goes positive, no current flows in the input grid circuit and therefore the input impedance is always infinite, giving the detector its name. In contrast to the anode-bend detector, the input impedance can be more truly said to be infinite because the effect of the grid-cathode capacitance is negligible since there is such a small voltage developed across it. As a result, the frequency response of this detector is excellent. Due to the negative feedback on the grid the a-f amplification closely approaches but cannot exceed 1 as for a diode detector. Another result of the negative feed back is that second harmonic distortion is considerably reduced for a-f.

The infinite-impedance detector compares favourably with the diode detector in the size of the applied signal voltage permissible and therefore in the amplitude of the output modulating-frequency voltage which it can provide. The output may approach half the anode supply voltage without overloading. For minimum distortion it is an advantage to apply a fairly large signal voltage. However, it falls short of the diode detector in that automatic volume control is not easily obtainable from the same circuit.

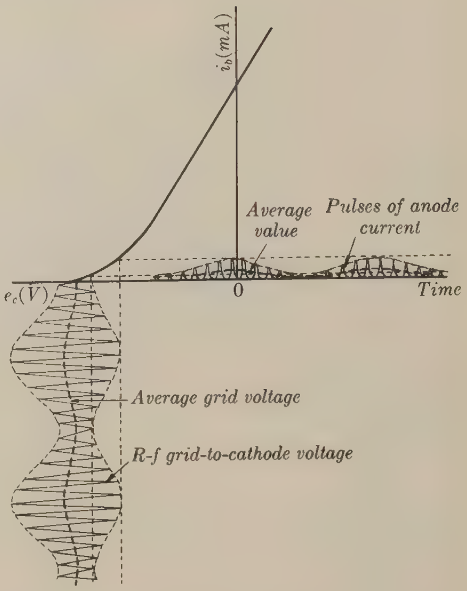


FIG.7-27 $e_c - i_b$ curve illustrating infinite-impedance detection.

This detector is becoming increasingly popular where a distortionless detector is needed for large signals and where, due to the demand of good selectivity, a current drain on the input circuit would be a particular disadvantage.

7.8 Vacuum-tube voltmeters. A *vacuum-tube voltmeter* is a widely used and important piece of radio test equipment. It consists essentially of a detector in which the change of anode current due to an input voltage can be calibrated in terms of the input voltage. It has several very decided advantages over the current-operated electromagnetic type of voltmeter, some of which are:

(i) a very high input impedance which may be as high as $100 M\Omega$ and seldom is lower than $10 M\Omega$

(ii) a wide frequency response; a well designed meter can be calibrated at 60 c/s and still give accurate readings at 50 Mc/s

(iii) high sensitivity; the voltmeter as a whole can be made very sensitive while using only a reasonably sensitive, and therefore cheaper, milliammeter in its circuit.

Types of Detectors and Their Characteristics

TYPE	LINEARITY	INPUT IMPEDENCE AND RESULTING SELECTIVITY	SENSITIVITY $= \frac{\text{A-F OUTPUT}}{\text{I-F INPUT}}$	FREQUENCY RESPONSE	SIGNAL-HANDLING CAPABILITIES	USES
Diode Detector	Very good on large signals	Low Poor selectivity	Poor Max. = 1	Excellent Interelectrode capacitance small	Excellent	1. High-level detector 2. High-frequency detector
Crystal Detector	Fair	Low Poor selectivity	Poor	Excellent	Fair	High-frequency detector
Weak-Signal Cumulative-Grid Detector or Grid-Leak Detector	Poor Square-law relationship	Low Poor selectivity	Very good Large amplification by valve	Poor Large input capacitance	Weak signals only	1. Where gain is more important than fidelity 2. Battery sets
Power Grid-Leak Detector	Fair	Low Poor selectivity	Very good	Poor	Strong signals only	In regenerative receivers
Lower-Anode-Bend Detector or Plate Detector	Fair	High Good selectivity	Good Amplification by valve	Poor	Fair	1. Inexpensive broadcast receivers 2. Amateur communications receivers
Upper-Anode-Bend Detector	Fair	Not as high as above	Good	Poor	Fair	
Infinite-Impedance Detector	Very good	Infinite Very Good selectivity	Poor Max. = 1	Excellent	Very Good	Where distortionless detection of large signals is required

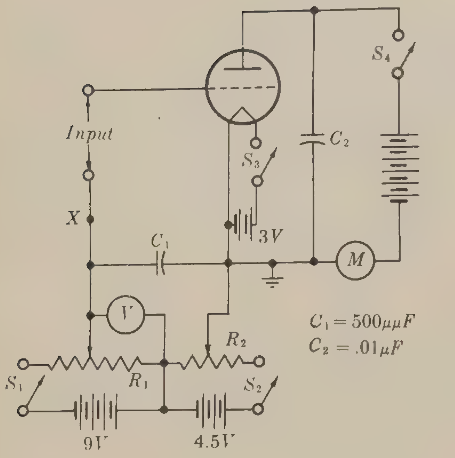


FIG. 7-28 Slide-back vacuum-tube voltmeter.

The vacuum-tube voltmeter is therefore particularly useful in circuits in which the flow of meter current appreciably alters the voltage to be measured as in r-f circuit measurements, measurement of amplifier output voltages, or of the voltage of a charged capacitor.

The vacuum-tube voltmeter may take many forms. Some measure a-c only, others d-c only and some both. Some measure r.m.s., some average and others the peak value of the applied voltage. In the following sections three of the most common types are discussed briefly.

A. Slide-Back Vacuum-Tube Voltmeter

This type of meter measures the peak value of the input voltage. A circuit is given in Fig. 7-28. The action is summarized as follows with the aid of Fig. 7-29. The triode acts as a lower-anode-bend detector. The grid bias is adjusted near to cut-off so that a very small anode current flows. The signal is then applied and the anode current flows in pulses and its average value increases. The grid bias is again increased until the average anode current equals its initial value. The increase in grid bias is taken to be the same as the peak voltage of the positive half cycle of the input signal.

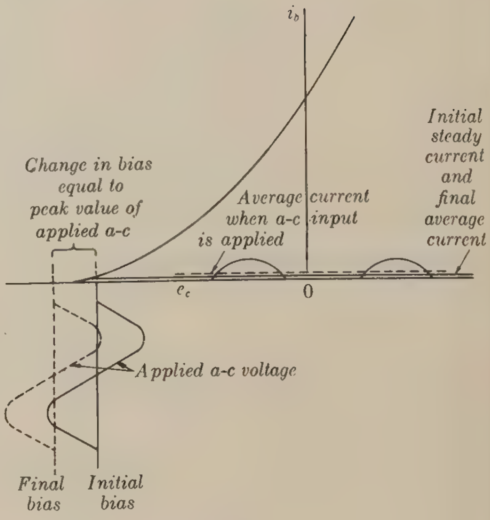

i_b
Initial steady current and final average current
Change in bias equal to peak value of applied a-c
Average current when a-c is applied
current input
e_c
0
Applied a-c voltage
Final bias
Initial bias

FIG. 7-29 $e_c - i_b$ curve illustrating operation of slide-back vacuum-tube voltmeter.

Actually this gives a value which is a little lower than the true peak voltage since the change in bias permits the peaks of the a-c to exceed the initial bias in order that the average bias they establish is the same as the initial bias and causes the same anode current to flow.

A more detailed consideration of the operation of the circuit and the function of its various components now follows. This is given in steps which are in the same order as is necessary for actual operation of the circuit.

(i) The meter is switched on. All the switches S_1, S_2, S_3, S_4 are ganged into one master switch.

(ii) R_1 is turned to the extreme right so that the voltmeter V reads zero. Then the input terminals are shorted. The potentiometer R_2, which adjusts the negative grid bias, is turned until the meter M in the anode circuit registers only a few microamperes. This is called the *false zero* reading.

(iii) The short across the input is removed and the unknown voltage applied to the input terminals. C_1 is a by-pass capacitor across the bias circuit so that any a-c input voltage appears only between grid and cathode. On the application of a signal the anode current rises. C_2 by-passes the H.T. supply for all a-c voltages.

(iv) R_1 is turned to the left to increase the grid bias until the anode current is reduced to the original false zero reading.

(v) The reading on the voltmeter is taken to be the peak value of the positive half cycles of the input voltage.

(vi) If the voltage to be measured is greater than 9 volts, additional bias can be placed at the point marked X. The input voltage is then the voltmeter reading plus the additional battery voltage.

Although the slide-back vacuum-tube voltmeter always gives a value which is a little lower than the true voltage this error becomes less if the applied voltage is large and the tube characteristic has a sharp cut-off. Another disadvantage of this type is that it is very difficult to use with fluctuating input voltages since the operator must have time to adjust the slide-back voltage. However, it has a very high input impedance since the grid draws no current and it can be used to measure very high a-c and d-c voltages. In addition no special calibration of the instrument is required.

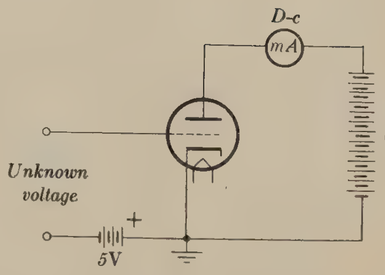


FIG.7-30 Vacuum-tube voltmeter employing lower-anode-bend rectification.

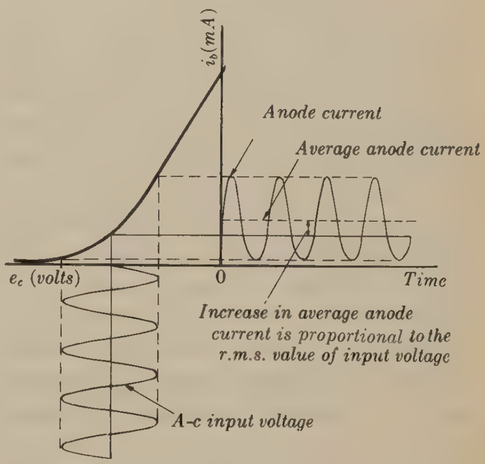

$i_b(mA)$
Anode current
Average anode current
e_c (volts)
0
Time
Increase in average anode current is proportional to the r.m.s. value of input voltage
A-c input voltage

FIG.7-31 $e_c - i_b$ curve illustrating full-wave square-law operation of voltmeter of Fig. 7-30.

B. Vacuum-Tube Voltmeter Employing Lower-Anode-Bend Rectification

Probably the most commonly used type of vacuum-tube voltmeter consists of a triode lower-anode-bend detector. Fig. 7-30 shows the basic circuit. The a-c voltage to be measured is applied between grid and cathode. The deflection of the d-c milliammeter needle is proportional to the r.m.s. value of the a-c input voltage. The milliammeter M is calibrated to read r.m.s. values of the input voltage directly. The characteristics of such a voltmeter depend on the adjustments. If the bias is adjusted to the middle of the curved region of the valve $e_c - i_b$ curve this is called *full-wave square-law operation* as in Fig. 7-31. The anode current is proportional to the r.m.s. value of the applied voltage, independent of the applied waveform. Since the grid is not biased to cut-off, a small anode current flows for zero input signal. A network consisting of a battery and rheostat across the meter is added to balance out this zero-signal anode current and thus permit the use of a more sensitive milliammeter (Fig. 7-32). In practice it is usually also necessary to provide a grid-leak resistor as shown to ensure a d-c path to the grid so that bias

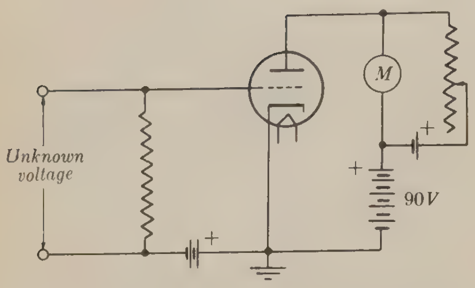


Fig.7-32 Voltmeter with circuit for balancing out zero-signal anode current.

is always applied. With this circuit the meter measures only one range of voltages, for example, 0.1-1.25V. If higher voltages are to be measured a different circuit arrangement must be employed. One common method used in practice is to convert the instrument to a peak-reading voltmeter by changing the circuit to a slide-back type by means of a range selector switch.

Instead of the full-wave square-law operation the circuit is frequently adjusted to provide *half-wave square-law operation.* This is accomplished by setting the grid bias exactly at cut-off as shown in Fig. 7-33. The anode current meter reads r.m.s. values as long as the signal voltage is small enough to allow square-law rectification. For larger signals the meter reading approximates average current values. With this adjustment it is not possible to calibrate the scale of the a-c milliammeter in terms of the a-c voltage applied to the grid except by experimental methods. However, once it is calibrated it is quite accurate for a-c voltages of any frequency up to about 30 Mc/s. The instrument may also be used for measuring d-c voltages as long as they are connected with the correct polarity, that is, positive to the grid and negative to the bias battery side. At no time must the grid be made positive with respect to the cathode since this allows it to draw current and therefore lower the input impedance of the meter.

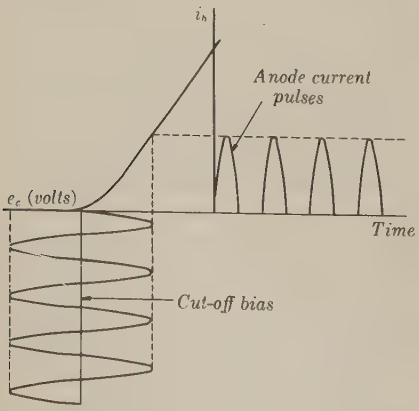


Fig.7-33 $e_c - i_b$ curve illustrating half-wave square-law operation of the voltmeter of Fig. 7-30.

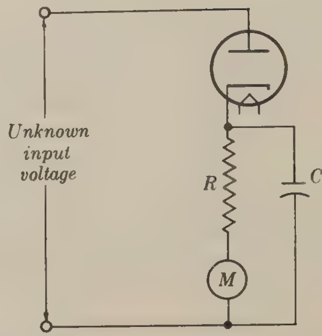


Fig.7-34 Diode vacuum-tube voltmeter.

C. Diode Vacuum-Tube Voltmeter

The simplest vacuum-tube voltmeter of all, consists of a linear, peak, diode detector with a meter to measure the voltage developed across C as shown in Fig. 7-34. The diode conducts on the positive half cycles of the input voltage charging C to almost the peak of the input voltage. The resistance R in series with the sensitive milliammeter together form a voltmeter which measures the voltage across C and thus the peak value of the input applied voltage. This is a useful circuit for the measurement of very high voltages when a slight flow of current is permissible. However, for small voltages, this circuit in itself is

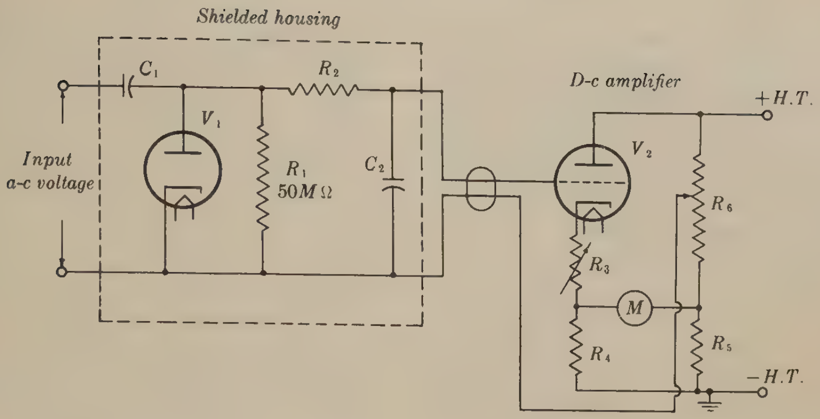


FIG. 7-35 Diode vacuum-tube voltmeter with d-c amplifier.

usually not sufficiently sensitive, consequently the diode detector is often followed by a d-c amplifier. The increase in current in the anode circuit of the d-c amplifier is proportional to the input voltage and is calibrated to read input voltage directly.

An actual commercial circuit of this type is shown in Fig. 7-35. V_1 and R_1C_1 form the diode detector. C_2R_2 filter out any a-c components appearing across R_1. R_3, R_4, R_5 and R_6 set the bias voltage for the d-c amplifier. R_6 varies the bias to balance out the zero-signal anode current so that the meter reads zero. The presence of R_3 causes negative feedback which permits the measurement of much higher voltages than it is possible to measure without it. R_3 is adjustable to provide several ranges of full-scale voltage. This type of meter is more sensitive than the anode-rectification type but the input impedance is lower, of the order of 10-15$M\Omega$. The components are specially designed and arranged to permit measurement at very high frequencies. At such frequencies, transit-time effects in the diode and series resonance of the input capacitance and inductance in the leads must be guarded against. This is accomplished by using an acorn valve and mounting V_1, R_1, C_1, R_2 and C_2 in a compact shielded housing with a short probe as the input lead. This rectifier unit is connected to the amplifier by means of a shielded cable. The net result of such construction is that the meter can be used at frequencies up to 100 Mc/s. The calibration is usually in terms of r.m.s. voltage although the meter actually registers the peak value of the input waveform. The peak value can be obtained by multiplying the r.m.s. reading by 1.414.

D. Importance of Waveform in Vacuum-Tube Voltmeter Measurements

In the preceding sections, reference is made in two ways to the various types of vacuum-tube voltmeters, one according to the type of rectifier used in the circuit, the other with respect to the response of the voltmeter. By the *response* of the voltmeter is meant whether the meter responds to peak, r.m.s. or average values of the input voltage. Each of these types of a-c voltage readings is needed for different applications. In practice vacuum-tube voltmeters are frequently not calibrated in accordance with the type of response characteristic of that particular meter. For example, a meter which actually responds to the peak value may be calibrated in terms of the r.m.s. value of the input voltage and vice-versa. This gives accurate readings as long as the relationship between the peak and the r.m.s. values of the input voltage is known. All vacuum-tube voltmeters are calibrated on the basis of a pure sine wave input. The relationships between peak, average and r.m.s. or effective values of a voltage having a pure sine waveform are known.

These relationships are as follows.

Peak voltage = r.m.s. voltage × 1.414

or r.m.s. voltage = peak voltage × .707.

Average of full-wave rectified sine wave = 63.6% of peak.

Average of half-wave rectified sine wave = 32% of peak.

However, if the input voltage has a complex waveform differing considerably from a pure sine wave, these relationships no longer hold. Some complex waveforms are shown in Fig. 7-36. In parts (a), (b) and (c) the average and r.m.s. values for a peak of 100*V* are also marked on the diagram. With such input waveforms it is possible for very serious errors to be made in voltage measurements. For example, suppose a peak-measuring instrument is calibrated to read r.m.s. values. In case (a) the peak voltage is 100*V*. On the voltmeter this is read as the r.m.s. voltage on the basis of a sine wave input i.e. 100 × .707 or 70.7 volts. Actually the r.m.s. voltage is 100*V*. A similar type of error can be made in an instrument which measures r.m.s. and is converted to peak values.

Another type of error is possible with peak-measuring instruments. It occurs when the positive peak value of the input voltage is different from the negative peak value as in Fig. 7-36 (d). This situation is detected by reversing the input terminals on the voltmeter when taking the measurement and noting any change in reading. For this reason it is sometimes referred to as *turnover*. The true r.m.s. value of such an input voltage is approximately equal to the average of the r.m.s. values obtained from each of the peak voltage readings.

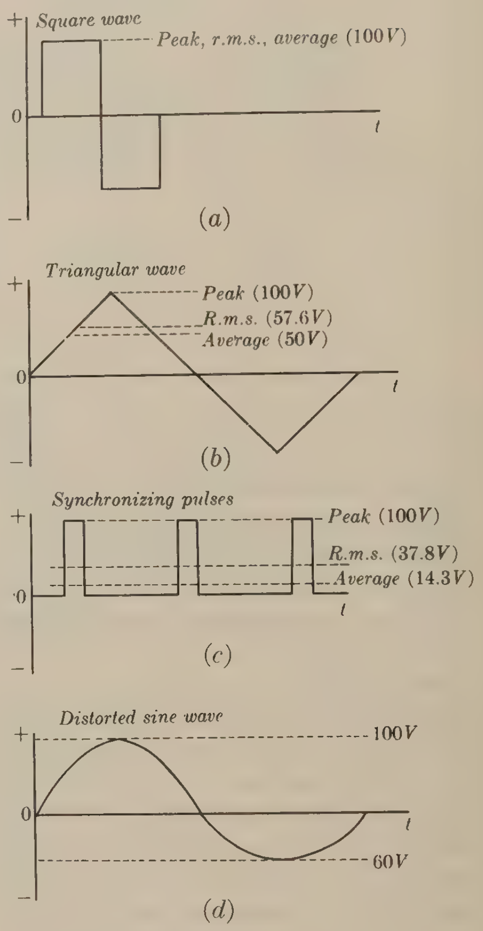


FIG. 7-36 Complex voltage waveforms.

Due to the possibility of turnover and because the r.m.s. value is a different percentage of the peak value for different complex waves, it is seen that it is very important to check the waveform of the input voltage when using a vacuum-tube voltmeter if any accuracy is to be obtained.

7.9 Applied mathematics.

A. Distortion in Diode Detectors

In the section on diode detection it is pointed out that a linear valve curve is not sufficient in itself to guarantee a distortionless a-f output. The size of the

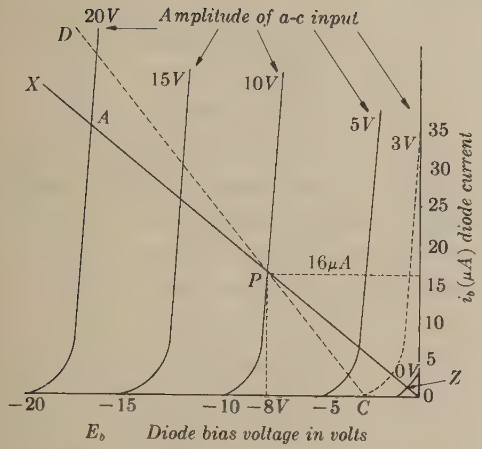


FIG.7-37 Dynamic curves for diode detector.

load presented to the a-f is an important consideration in determining whether distortion takes place or not. In order to understand the reason for this, it is necessary to refer to the dynamic curves of a diode (Fig. 7-37). These are obtained experimentally by means of a circuit of the type shown in Fig. 7-38. The amplitude E of input alternating voltage is held constant while the bias voltage E_b is varied. E_b represents the d-c voltage which is normally developed across the load resistor and which applies a negative bias to the anode of the diode. For each given amplitude of alternating voltage, the rectified current values, measured by the microammeter M are plotted against the corresponding values of E_b.

Now suppose the diode load resistor R is .5$M\Omega$ (Fig. 7-38). In Fig. 7-37 OX represents the d-c load line, which is merely a plot of voltage across R against current through R. If the input carrier voltage is 10V peak, the intersection of the dynamic curve for a 10V input and the load line OX at the point P determines that the current is 16μA while the d-c voltage is $-8V$. If the carrier is 100% modulated, the operating point moves at an a-f rate along the load line from Z at the intersection with the zero-signal curve, through P, to A at the intersection with the 20V curve and back again. The changes in current and voltage across R over this region are almost exactly proportional to the input modulation envelope. A small amount of distortion, 5% or less, is produced in the region near the zero-signal curve, which may be seen from the fact that PZ is not the same length as PA.

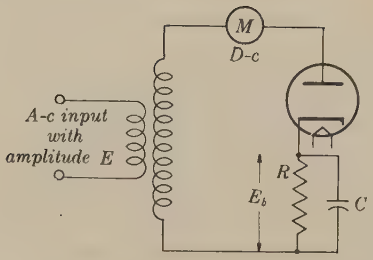


FIG.7-38 Circuit for obtaining dynamic curves of diode detector.

However, if the diode load is shunted by the input circuit of the next stage, the total effective a-c load for a-f is considerably reduced. Assuming that the total effective a-c load is 250,000Ω the a-c load line is given by the dotted line CD. Although it has a new slope it still passes through the operating point P, because for no modulation the conditions for the d-c load of .5$M\Omega$ still hold. This new load line reaches cut-off when the r-f signal is in the neighbourhood of 3V peak. Therefore, if 100% modulation is used, the negative peaks are flattened considerably creating a high percentage distortion. For this case the input signal would have to be limited to 70% modulation in order to overcome this effect. The only other alternative to limiting the percentage modulation, in order to prevent distortion, is to ensure that the impedance represented by the a-c load line CD is the same as the d-c impedance given by the load line OX. This is usually accomplished by making the value of the grid-leak resistor of the following stage three or four times the size of the diode load resistor.

B. *Mathematical Analysis of Square-Law Detection*

This analysis holds only for small signal voltages which operate over the curved portion of the valve current-voltage curve (Fig. 7-24). This small region at the bottom of the curve is approximately a parabola and therefore can be represented by an equation of the form $i_b = p+qe_c+re_c^2$ where p, q and r are constants whose values depend on the curvature of the particular dynamic curve in question. Let the steady zero-signal voltage on the grid be E_c and the steady anode current is then given by $I_b = p+qE_c+rE_c^2$. Now let an alternating signal voltage $E_s \sin \omega t$ be applied to the grid. Since this is superimposed on the steady voltage E_c, the resulting grid potential e_c is given by $E_c+E_s \sin \omega t$. The corresponding current i_b is then equal to

$$p+q(E_c+E_s \sin \omega t)+r(E_c+E_s \sin \omega t)^2$$
$$=p+qE_c+qE_s \sin \omega t+rE_c^2+2rE_cE_s \sin \omega t$$
$$+rE_s^2 \sin^2 \omega t$$

From trigonometry

$$\cos 2\omega t = 1-2 \sin^2 \omega t$$

or $$\sin^2 \omega t = \tfrac{1}{2}(1-\cos 2\omega t)$$

Therefore i_b is

$$=p+qE_c+qE_s \sin \omega t+rE_c^2+2rE_cE_s \sin \omega t$$
$$+\frac{rE_s^2}{2}(1-\cos 2\omega t)$$

and from above

$$I_b = p+qE_c+rE_c^2$$

Therefore i_b becomes

$$i_b = I_b+\frac{rE_s^2}{2}+(qE_s+2rE_cE_s)\sin \omega t$$
$$-\frac{rE_s^2}{2}\cos 2\omega t.$$

The last two terms represent radio-frequency currents for which the average value over one complete cycle or at least over several cycles is zero. This is another way of saying that they are by-passed by the r-f by-pass capacitor in the filter circuit. The term $\frac{rE_s^2}{2}$ represents the change in the average value of anode current. This shows that the rectified current is proportional to the square of the amplitude of the applied voltage.

Also $$i_b = p+qe_c+re_c^2$$

Therefore, $$\frac{di_b}{de_c} = q+2re_c$$

and $$\frac{d^2i_b}{de_c^2} = 2r.$$

Hence the change in the average anode current, $\frac{rE_s^2}{2}$, may be written in the form

$$\frac{E_s^2}{4}\times 2r$$

or $$\frac{E_s^2}{4}\times\frac{d^2i_b}{de_c^2}$$

This shows that the rectified current is also proportional to the rate of change of slope of the e_c-i_b curve. Therefore, the most sensitive region for rectification occurs about the point where the slope is changing most rapidly, i.e. at the upper and lower bends of the curve.

2.10 Questions and problems.

1. What is detection or demodulation?
2. Why is detection necessary?
3. What are the two fundamental steps in detection?
4. Draw a simple diode detector circuit and explain its operation, using waveforms to illustrate.
5. Draw a simple full-wave diode detector circuit.
6. Why is less filtering required with a full-wave diode detector circuit than with a half-wave diode detector circuit?
7. What is the disadvantage of a full-wave diode detector?

8. In a simple diode detector the value of the load resistance is .5$M\Omega$. Calculate the value of the by-pass capacitor if the audio frequency involved is less than 7000 c/s.

9. In the filter circuit in a diode detector, if $R = 1M\Omega$ calculate the value of C. Check your value by showing that X_C is small for r-f and large for a-f currents.

10. In question 9, what type of distortion results if C is chosen too large?

11. What is meant by the rectifying ratio of a crystal?

12. How is the rectifying ratio of a crystal seriously reduced?

13. Draw a circuit for a crystal detector and explain its action.

14. What particular features of crystal detectors are mainly responsible for their again coming into use?

15. What is meant by square-law detection? Why is weak-signal detection a suitable alternative name?

16. Draw a circuit diagram of a weak-signal cumulative-grid detector and explain its operation using waveform diagrams.

17. Describe any changes in the circuit of question 16 in order that it may act as a power grid-leak detector and describe the operation.

18. Draw a circuit for a lower-anode-bend detector and explain its action using diagrams.

19. Describe the changes necessary to convert the circuit of question 18 into an upper-anode-bend detector and explain the operation.

20. Why are only pentodes used with upper-bend detection?

21. Explain why distortion takes place when a very large signal is applied to a linear anode-bend detector.

22. What is the effect of the grid-to-cathode capacitance in an anode-bend detector?

23. Draw the circuit for an infinite-impedance detector and explain its operation with the aid of diagrams.

24. Why is the input impedance of an infinite-impedance detector even higher than that of an anode-bend detector when neither draws current?

25. Name five types of detectors and compare them under the following headings
    (a) linearity
    (b) selectivity
    (c) sensitivity
    (d) signal-handling capabilities.

26. On what factors does the sensitivity of a detector depend?

27. Draw a circuit for a simple slide-back vacuum-tube voltmeter. Explain the operation giving the purpose of each circuit component.

28. Name three types of vacuum-tube voltmeters and give advantages and disadvantages for each.

29. Explain the meaning of waveform error in connection with vacuum-tube voltmeter measurements.

CHAPTER VIII

RADIO COMMUNICATION TRANSMISSION

8.1 Principles of transmission. The limitations of transmission by means of sound waves are discussed in an earlier chapter. A material medium, solid, liquid or gaseous is required for the transmission of sound. Sound waves in air are quickly attenuated so that long distance communication by this means is not practicable. Radio waves, on the other hand, do not require any material medium for propagation. They will travel over very great distances with small attenuation. A radio transmitter and its aerial system are used to produce radio waves which travel from the transmitter to the receiver. The intelligence which is to be communicated is superimposed upon the radio wave by the process of modulation which takes place in the transmitter. The receiver recovers the intelligence from the radio wave at the receiving station.

8.2 Essentials of a transmitter. The essential components of a radio transmitter are shown in the block diagram of Fig. 8-1. The *oscillator* generates a high-frequency alternating current which is called the r-f carrier. The oscillator operates at a low power level in order to maintain an output of constant frequency. This r-f current from the oscillator is amplified by the *power amplifier* and then passed to the transmitting aerial. Several stages of amplification may be used in the power amplifier section. The high-frequency current flowing in the aerial wire sets up radio waves which travel out from the transmitter.

The *modulator* superimposes the intelligence to be communicated upon the r-f carrier. The modulator consists of a microphone and audio-frequency amplifier for r-t transmission. This circuit is replaced by a tone generator if m-c-w transmission is to be used. When c-w transmission is employed, the modulator is not required and a key is used to interrupt the r-f carrier.

The *power supply* provides the H.T. and L.T. voltages required for operating the circuits in the oscillator, power amplifier and modulator section. If the transmitter is a low-power one, a single power supply may be used for all stages. If the transmitter is one of high power, separate supplies are usually employed for the oscillator and power amplifier circuits. A low-voltage, well regulated supply is used for the oscillator in order to avoid changes in frequency due to changes in supply voltage.

The waveforms of the output of each section are shown in the block diagram of Fig. 8-1 during the transmission of m-c-w signals.

The main consideration in transmitter design is the necessity for *frequency stability*. This may be defined as the ability of a transmitter to transmit at a constant frequency during the total time of transmission. In any frequency band each transmitter must maintain a stable

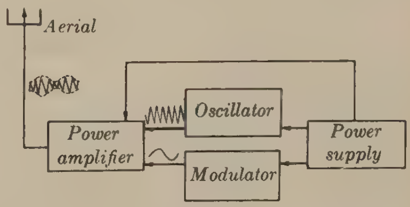


FIG.8-1 Block diagram of a transmitter.

frequency in order to prevent interference at the receiver caused by stations operating on adjacent channels. In addition, the transmitter must be stable in frequency so that the receiver will not have to be continually adjusted to be kept in tune with the transmitter. The difficulty of keeping tuned to a weak signal which is unstable in frequency might be great enough to prevent the operator from copying the message. Also, while maintaining a listening watch on some frequency, the operator might even miss the signal if it were not on frequency. Thus good reception depends to a large extent upon the frequency stability of the transmitter.

To eliminate the difficulties mentioned above, the frequency tolerances for different types of transmitters in the various frequency bands have been laid down as international standards. As an example, the table at the bottom of this page gives the frequency tolerances for transmitters in the frequency band from 10 to 550 Kc/s.

The above frequency tolerances state the maximum permissible separation between the actual frequency being transmitted and the assigned frequency as a percentage of the latter. For ship stations, the reference frequency is the frequency on which the transmission is first observed and the figure in the table marked by an asterisk refers only to frequency separations permitted during the ten-minute period of transmission following first observation. The frequency tolerances given above do not take into account sidebands caused by modulation.

8.3 Frequency stability.

A. Frequency Drift

The term *frequency drift* refers to the gradual change in frequency of the oscillator caused by a change in the temperature of the components. In the circuits following the r-f oscillator the output frequency is the same as the input frequency except where a multiplier (section 8.8) is used. In this case the output frequency is some exact multiple of the input. Frequency stability in oscillators is also discussed in Chapter V.

Changes in temperature of the oscillator components may be caused by changes in loading of the oscillator, gradual heating of the valve, changes in external temperature and changes in electrode or filament voltages. Usually an increase in temperature causes an increase in the inductance and capacitance so that the resonant frequency of the tuned circuit decreases.

Changes in frequency due to external temperature changes may be avoided by placing the oscillator in a compartment whose temperature is controlled.

One way to compensate for changes in the circuit values is to use a negative-temperature-coefficient capacitor. As the

TYPES OF TRANSMITTING STATIONS	TOLERANCE
(i) Fixed Stations	±0.1%
(ii) Land Stations	±0.1%
(iii) Mobile Stations using frequencies other than those of bands indicated under (iv)	±0.1%
(iv) Mobile Stations using frequencies of the bands: 110 to 160 Kc/s and 365 to 515 Kc/s	±0.3%*
(v) Aircraft Stations	±0.3%
(vi) Broadcasting	±20 c/s

temperature rises the capacitance of the capacitor decreases, offsetting the change in size of the other components. The negative temperature coefficient of capacitance is specified in $\mu\mu F/\mu\mu F/{}^{0}C$ and ranges from 0.0 to $0.001\mu\mu F/\mu\mu F/{}^{0}C$. The choice of the proper capacitor with the proper temperature coefficient depends upon the circuit and the one compensating component can eliminate frequency drift arising from all sources.

The use of specially constructed anti-drift coils is another method of compensating for frequency drift. If a coil is wound on a form having a low temperature coefficient of expansion, the change in dimensions is minimized. When a flat conductor is wound on this type of form, the change in inductance can be reduced to 5-$15\mu H/H/{}^{0}C$. The coil itself may be painted black so that the radiation of heat takes place more readily and the temperature does not rise as high. Another type of anti-drift coil is wound on a former with a longitudinal slot. Each turn of the coil is wound as in Fig. 8-2 to allow the coil and former to expand and contract without an appreciable change in diameter and hence in inductance.

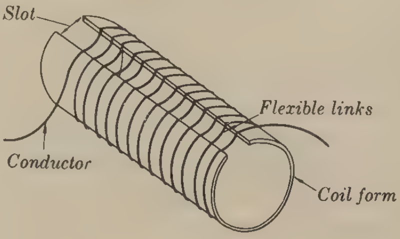


FIG.8-2 Anti-drift oscillator coil.

B. Q of Tank Circuit

The frequency at which an oscillator operates is not exactly the resonant frequency of the oscillatory circuit. The frequency is such that the phase and amplitude of the voltage fed back to the grid are correct to maintain oscillation. This is very close to the resonant frequency of the tank circuit and depends upon the load on the circuit, the harmonics generated, the stray reactances, the voltages applied to the valve and the Q of the tank circuit.

The amount by which the operating frequency of the oscillator differs from the resonant frequency of the tank circuit is inversely proportional to the effective Q of the tank circuit. The higher the Q the smaller is the change in frequency which can be caused by the other factors. A high Q is obtained by using loose coupling for the feedback to the grid circuit and to the output, and by using a low L/C ratio in the tank. However, loose coupling to the load results in lower power output. Thus, the Q is usually set at about 12 for efficient operation.

C. Effect of Loading

When the oscillator is loaded, an impedance is reflected back into the tank circuit which may change its resonant frequency and hence the operating frequency. The resonant frequency of the circuit depends chiefly on the values of L and C and only slightly on the value of R. Thus if the load is highly inductive or capacitive the change in the resonant frequency, and therefore in the operating frequency, is great. In addition, a greater amount of loading causes a greater change in operating frequency. For best stability the coupling should be as loose as possible to prevent excessive loading and the reflected load should be a pure resistance. A *buffer amplifier* is usually employed between the oscillator and the output load to reduce the loading on the oscillator. The oscillator can then be operated well below its maximum power-handling capabilities. A low-power oscillator can be used because of the power amplification provided by the buffer.

D. Supply Voltage

Variations in supply voltages cause frequency variations by changing the valve factors which affect the components of the oscillatory circuit. The interelectrode capacitances depend to a certain extent on the space charge and the electron density in the valve. Both of these are affected by a change in anode voltage, grid voltage, or filament voltage. To eliminate this source of frequency variation, regulated anode, screen and filament voltage supplies must be used.

E. Harmonics

Harmonics generated in the oscillator are detrimental to its frequency stability. The heterodyning (section 9.5) of these harmonics produces fundamental frequency components which are not necessarily in phase with the fundamental frequency currents. This will change the resultant phase of the fundamental and cause a frequency change. The use of a high-Q tank circuit minimizes this effect.

F. Other Causes of Frequency Instability

Frequency instability in the oscillator is also caused by:

(i) mechanical vibration of components such as capacitor plates, lead wires and coil turns
(ii) feedback of energy from the output circuit to the input which is not in the correct phase or which has not the correct amplitude
(iii) parasitic oscillations caused by stray inductance and capacitance
(iv) keying of the oscillator circuit.

The capacitance of a variable tuning capacitor may change considerably with vibration and so cause frequency variations. To avoid the use of such a capacitor, a variable inductor is sometimes used along with a fixed capacitor whose capacitance does not change with vibration. Rubber mountings and rigid connecting wires are also used to reduce mechanical vibration to a minimum. The elimination of unwanted feedback through stray inductive and capacitive coupling is accomplished by proper shielding between stages. Neutralization (section 9.9) counteracts the effect of the interelectrode capacitances. Parasitic oscillations and undesirable feedback are thus reduced. To prevent frequency variation caused by sudden stopping and starting of the oscillator when it is being keyed, it is usually operated continuously and the stages following it are keyed.

8.4 Master-oscillator control. When the oscillator which generates the carrier frequency of the transmitter is not a crystal oscillator and is followed by one or more power amplifiers it is called a *master oscillator* (m.o.). The oscillator is designed for good frequency stability and operates at a low power level. The power amplifiers are well shielded and neutralized so that the unwanted oscillations do not occur. A frequency stability of 100 parts in a million can readily be obtained on low and medium frequencies. As the frequency is increased this stability becomes more difficult to attain. Master-oscillator control permits easy tuning of the transmitter to any frequency in a certain range. This *flexibility* is its main advantage.

8.5 Crystal control. The use of a crystal oscillator to control the frequency of the transmitter provides higher frequency stability than is obtainable from a master oscillator. Since the crystal may be damaged if it is over-driven, the power output of the oscillator must be kept to a very low value. This necessitates the use of more amplification after the oscillator than in the case of m.o. control. However the advantage of higher stability makes up for this disadvantage.

The danger of damaging the crystal

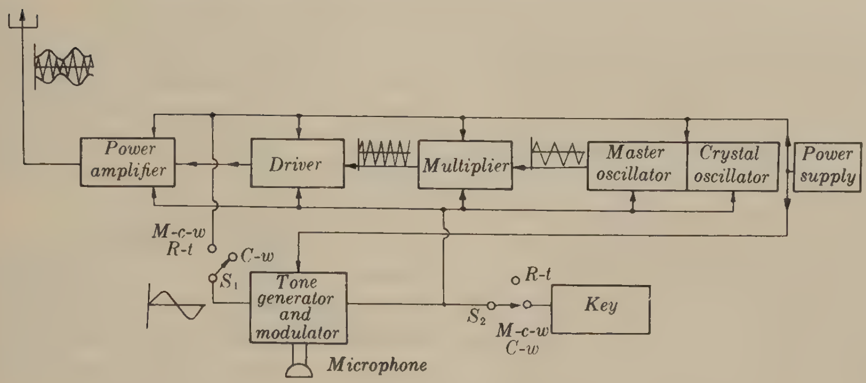


FIG.8-3 Block diagram of high-frequency high-power transmitter.

becomes greater as the frequency increases since thinner, more fragile crystals are used. There is therefore a high-frequency limit, above which a crystal cannot be made to operate. The use of a frequency multiplier or multipliers is then necessary in order to produce very-high-frequency carriers. Consequently a crystal-controlled transmitter usually has frequency multipliers and/or buffer amplifiers between the oscillator and the final power amplifying stage.

In circuits employing crystal control, a separate crystal is supplied for each carrier frequency. Hence the flexibility of a crystal oscillator is poor. In many cases a combination of crystal and m.o. control is used to make available the advantages of each, one or the other being used at any time.

8.6 Comparison of master-oscillator and crystal control.

MASTER-OSCILLATOR CONTROL	CRYSTAL CONTROL
Carrier frequency is determined by the tuning of the master oscillator.	Carrier frequency is determined by the resonant frequency of the crystal.
Low power is used to avoid frequency drift.	Low power is used to avoid frequency drift and damage to the crystal.
Amplifiers are necessary to increase the power.	Amplifiers are necessary to increase the power.
Relatively poor frequency stability	Very good frequency stability
Good flexibility	Poor flexibility

8.7 Block diagram of a high-frequency high-power transmitter. The transmitter illustrated in Fig. 8-3 is a high-frequency high-power transmitter for r-t and c-w transmission. The frequency-controlling stage may be either a crystal oscillator or a master oscillator. The crystal oscillator is used for good frequency stability and when it is used the master oscillator is connected as an amplifier. When using the master oscillator to generate the carrier frequency the crystal oscillator is inoperative. This combination allows complete coverage of the frequency band.

The *multiplier* stage is usually designed to act as an amplifier on one range and a frequency doubler or tripler on others. By this means the upper frequency limit can be extended to three times that of the master oscillator. The master oscillator range is so chosen that its upper frequency limit is double its lower frequency limit on each band. In this way continuous coverage may be obtained over all bands.

The output of the multiplier is higher in frequency but lower in power than its output when acting as an amplifier. Consequently an additional stage of power amplification is necessary to provide the large input needed to drive the power amplifier. This stage preceding the power amplifier is called the *driver.*

For c-w transmission, switch S_2 is closed and S_1 is open. The key controls all stages except the modulator which is not connected into the circuit. The keying is affected by applying and removing a large grid bias on the three stages of the transmitter. The bias is provided by a bleeder across the power supply which is allowed to operate continuously. When the key is open, bias is applied to cut off all current through the transmitter valves so that they do not operate. When the key is closed it shorts out the bias and allows the transmitter to operate normally. The name *paralyzing bias* is given to this bias since it paralyzes the transmitter when the key is open.

For m-c-w transmission the switch S_1 is also closed. The *tone-generator* modulator operates as an audio-frequency oscillator which generates a sinusoidal output at an audio frequency. This output is used to modulate the r-f carrier by grid or anode modulation of the power amplifier stage. Keying is accomplished as in the case of c-w transmission. A paralyzing bias is applied to the grids of all stages when the key is released. This bias is shorted to permit operation of the valves when the key is depressed.

For r-t transmission, switch S_1 is closed as in the case of m-c-w transmission and S_2 is opened to disconnect the key. In addition, the feedback circuit, which causes the modulator to operate as an audio-frequency oscillator on m-c-w, is disconnected. The modulator now operates as an amplifier to amplify the a-f signal from the microphone. The output of the modulator is used to modulate the r-f carrier in the power amplifier stage.

8.8 Frequency multipliers. The frequency multiplier is a tuned amplifier having its output circuit tuned to a harmonic of the input. The harmonic output is obtained by taking advantage of the fact that the pulses of anode current in a Class C amplifier have a high harmonic content. The output circuit is tuned to the desired harmonic of the fundamental and the amount of power produced in the tank circuit is greatest at this frequency because the impedance at this frequency is greatest. To all the other components of the anode current the impedance of the tank circuit is small so that the amount of power produced at these frequencies is small. Consequently the circulating current in the output tank circuit has a sinusoidal waveform and has the frequency of the harmonic to which it is tuned.

The power output from a multiplier decreases as the order of the harmonic increases. A multiplier whose output circuit is tuned to the second harmonic produces about 60 to 70 percent as much output power as is obtained when the tank circuit is tuned to the fundamental. For the third, fourth and fifth harmonics the percentages are approximately 40, 30 and 25 respectively. The circuit of Fig. 8-4 illustrates a frequency multiplier. The valve is biased well beyond cut-off for Class C operation. Hence, current flows through the valve for only a short time during the positive half of each input cycle when the signal drives the grid above cut-off. If the circuit is to be used as a frequency doubler the anode circuit, L_2C_2, is tuned to resonate at a frequency which is twice the frequency

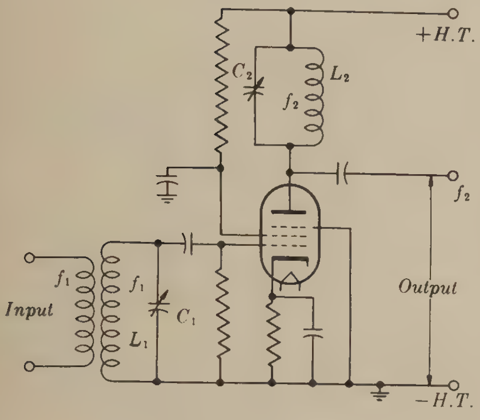


FIG. 8-4 Frequency multiplier.

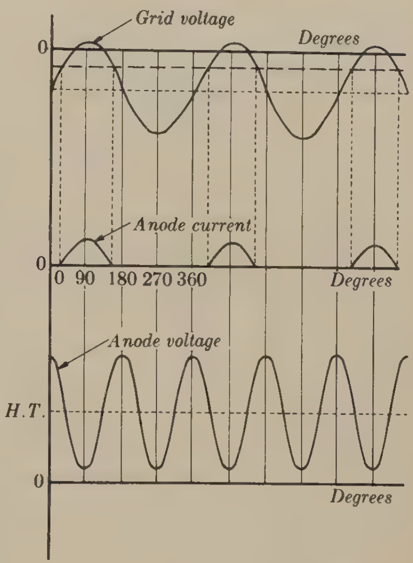


FIG. 8-5 Voltage and current waveforms in a frequency doubler.

of the input signal. Therefore two complete oscillations take place in L_2C_2 during each oscillation in L_1C_1 (Fig. 8-5). One pulse of current flows through the valve for every second oscillation in L_2C_2. These current pulses make up for losses in the circuit.

If the circuit is to be used as a frequency tripler, L_2C_2 is tuned to a frequency 3 times that of the signal. In this case, one pulse of current flows through the valve for every third oscillation in L_2C_2.

The efficiency of the circuit is determined by the operating angle used in the design. For optimum performance the angle of current flow is about $\frac{180}{n}$ degrees where n is the order of the harmonic. For a doubler this operating angle should be about 90° of the fundamental which is equivalent to about 180° of the second harmonic. Consequently in order to provide for such a small operating angle the bias must be considerably increased over normal Class C operation. To drive the valve to the same peak current as that obtained when amplifying at the fundamental frequency, the input voltage must be very much larger. Even then, the average d-c anode current is only a fraction of the value attained at the fundamental because of the smaller operating angle. Thus when the valve is operated as a doubler at its optimum angle the power output obtainable is only about 70% of that obtainable from a Class C amplifier in which input and output frequencies are the same.

Greater output can be obtained by using a large operating angle. This decreases the overall efficiency but the higher power output more than compensates for the reduced efficiency. A lower grid bias or a valve with lower anode resistance is used so that the anode current is increased. In doing this, the anode dissipation and grid dissipation in the valve must not be exceeded. The curves in Fig. 8-5 show the instantaneous current and voltage relations of a frequency doubler.

In the frequency multiplier, the Q of the tank circuit must be made large. This is necessary to minimize power losses in the tank circuit as the anode current flows for only a short time during each cycle. This is especially important in a multiplier because the current flows only once for every two cycles of the output frequency. The high Q also tends to eliminate the components of output at other frequencies which are present in the anode current. The grid impedance is high because the bias is large. This causes low power loss in the grid circuit. Neutralization is not necessary in a multiplier since the anode and grid circuits are not tuned to the same frequency. Thus the effects of feedback are not enough to cause oscillation.

A push-pull stage cannot be used as a frequency doubler because even harmonics are cancelled out by the push-pull operation. Since odd harmonics are produced, push-pull amplifiers can be used as frequency triplers. The same considerations obtain as in single-valve multipliers. For frequency doubling, a modified push-pull circuit is sometimes used as in Fig. 8-6. This circuit produces a fundamental frequency output which is twice the input frequency. The valves are biased so that they conduct on alternate half cycles of the input. When V_1 conducts, a pulse of current flows through the tuned output circuit producing one cycle of oscillation in L_2C_2. When V_2 conducts and V_1 is cut off, another pulse of current flows through the tuned circuit and produces a second complete oscillation in L_2C_2. Therefore two cycles of oscillation are produced in the anode circuit for each cycle of input. Waveforms of voltage input and current through the valves are shown in Fig. 8-7.

8.9 Neutralization. The feedback from the anode circuit to the grid circuit through the anode-to-grid interelectrode capacitance of a power amplifier valve has a serious effect on the frequency stability of the circuit as well as causing oscillations to occur. In tetrodes and pentodes the interelectrode capacitance is reduced by the screen grid. These valves are useful in applications requiring

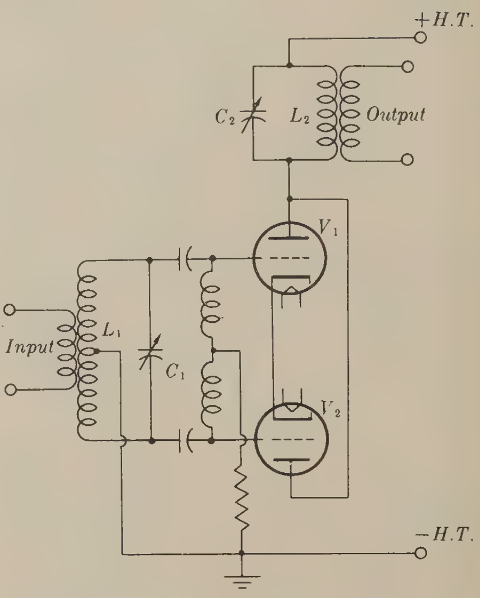


FIG.8-6 Two-valve frequency doubler.

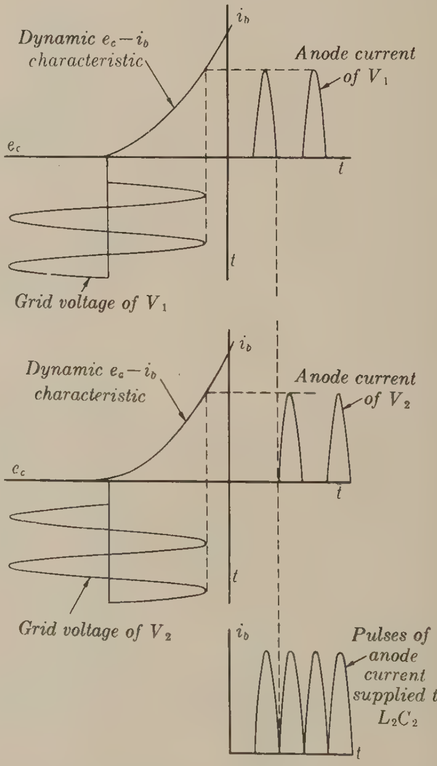


FIG.8-7 Operation of two-valve frequency doubler.

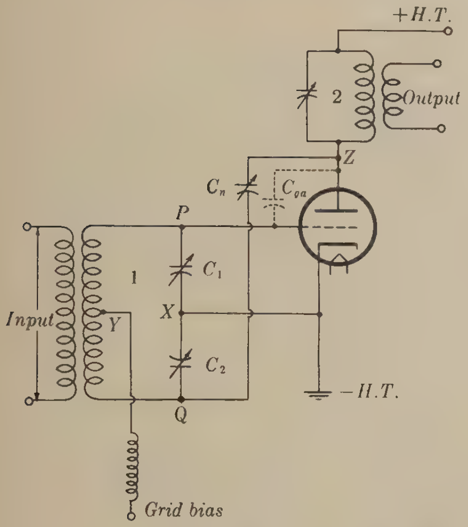


FIG. 8-8 Grid-neutralized amplifier.

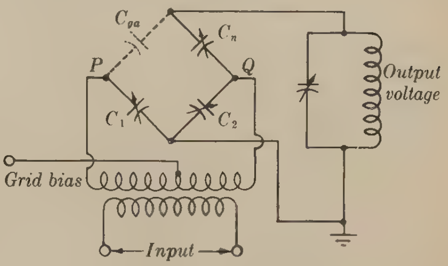


FIG. 8-9 Equivalent bridge circuit of grid neutralization.

comparatively low power output. However in the final ouput stage of a transmitter where large power must be produced, tetrodes and pentodes are not suitable and triodes are used. Since the anode-to-grid interelectrode capacitance cannot be eliminated in a triode, special circuit design providing *neutralization* must be used to prevent oscillation and frequency drift.

To neutralize a triode amplifier some of the r-f voltage must be taken from the output and introduced into the input circuit in such a way that this voltage effectively cancels that being passed through the anode-to-grid interelectrode capacitance. For complete neutralization, the two voltages must be opposite in phase and equal in amplitude. Circuits in which the out-of-phase voltage is taken from the grid circuit and fed to the anode circuit are called *grid-neutralized* circuits, while those in which the neutralizing voltage is fed from the anode circuit to the grid circuit are said to be *anode-neutralized.*

A common circuit for grid neutralization called the *Rice* neutralizing circuit is shown in Fig. 8-8. The object is to prevent energy in circuit (1) from being transferred to circuit (2) through the interelectrode capacitance, C_{ga}. The input circuit (1) is centre-tapped and balanced to ground at X and Y. When an r-f signal is applied to the primary circuit of the input transformer the potentials at P and Q in the secondary are in anti-phase. P and Q in the input circuit are capacitively coupled to the point Z in the output circuit. The coupling from P to Z is through the interelectrode capacitance C_{ga} while the coupling from Q to Z is through the neutralizing capacitor, C_n. If X is electrically midway between P and Q and C_n is adjusted to be equal to C_{ga}, the two paths have equal impedances and, since P and Q are in anti-phase, as much energy is passed toward Z along one path as is passed back from Z along the other. That is, changes in potential at Z due to changes in potential at P are counteracted by changes in potential at Z due to exactly opposite changes in potential at Q.

The action of this grid neutralizing circuit is also explained with the aid of the equivalent bridge circuit shown in Fig. 8-9. The input split-stator capacitor, grounded at the centre, consists of equal capacitances C_1 and C_2. The output voltage is divided across C_n and C_2 in one arm and C_{ga} and C_1 in the other arm. For a balance in the bridge, that is, no voltage difference between the points P and Q, the voltage across C_n

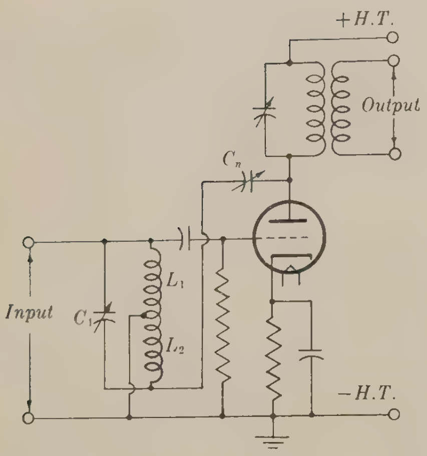


FIG.8-10 Modified grid neutralization.

and C_{ga} must be equal. But the voltage division across C_n and C_2 is inversely proportional to their capacitances.

Thus, $$\frac{E_{Cn}}{E_{C2}} = \frac{C_2}{C_n}$$

$$\therefore \quad E_{Cn} = \frac{E_{C2} \cdot C_2}{C_n}$$

Also, $$\frac{E_{Cga}}{E_{C1}} = \frac{C_1}{C_{ga}}$$

$$\therefore \quad E_{Cga} = \frac{E_{C1} \cdot C_1}{C_{ga}}$$

For a balance $E_{Cn} = E_{Cga}$

Therefore, $$\frac{E_{C2} \cdot C_2}{C_n} = \frac{E_{C1} \cdot C_1}{C_{ga}}$$

But $E_{C2} = E_{C1}$ and $C_2 = C_1$.

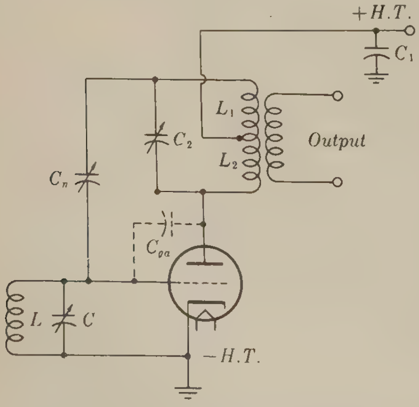


FIG.8-12 Anode-neutralized amplifier.

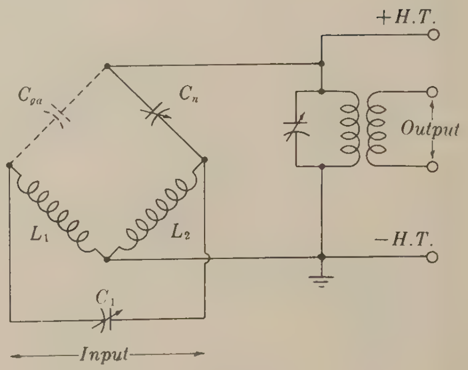


FIG.8-11 Equivalent bridge circuit of Fig. 8-10.

Therefore $$\frac{1}{C_n} = \frac{1}{C_{ga}}$$

Therefore $C_n = C_{ga}$.

Thus, to neutralize the circuit, C_n is adjusted to be equal to the anode-grid capacitance in the valve and the stray anode-grid capacitance outside the valve.

A modification of the grid neutralization circuit is shown in Fig. 8-10. The centre of the coil is grounded to provide anti-phase voltages at opposite ends of the tuned input circuit. A single input capacitor C_1 is used instead of the split-stator capacitor. The equivalent bridge circuit is shown in Fig. 8-11. The operation of the bridge circuit is the same as in the first case. As before, the condition for balance is that $C_{ga} = C_n$.

A simple anode-neutralized amplifier

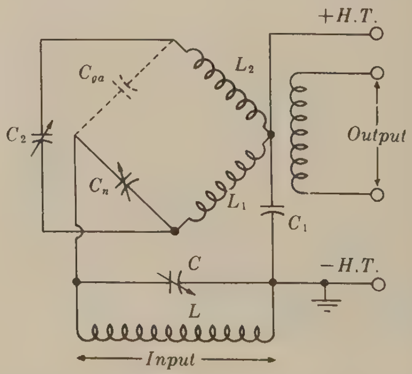


FIG.8-13 Equivalent bridge circuit of anode neutralization.

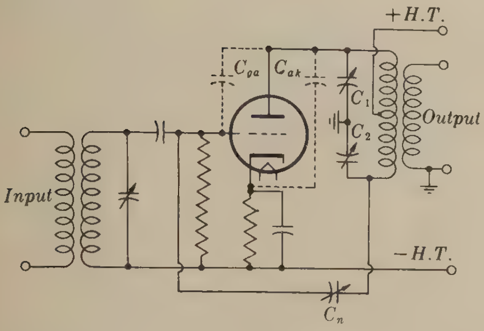


FIG. 8-14 Neutralized amplifier with split-stator capacitor.

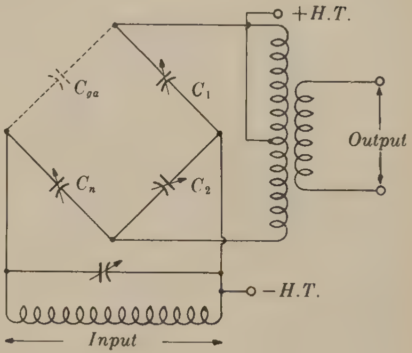


FIG. 8-15 Equivalent bridge circuit of Fig. 8-14.

is shown in Fig. 8-12. This circuit is known as the *Hazeltine* neutralizing circuit or the *neutrodyne* circuit. The tuned anode circuit is balanced to ground for a-c voltages by the centre-tap connection to the H.T. supply. This point is maintained at r-f ground potential by the by-pass capacitor C_1. The neutralizing capacitor C_n is approximately equal to C_{ga} and is connected to the end of the tank circuit opposite that to which the anode is connected. Hence the voltages at the ends of the coil are 180° out of phase with each other. Therefore as much energy is passed to the grid circuit through C_{ga} as is passed back to the anode circuit through C_n. The equivalent bridge circuit which must be balanced by adjusting C_n is shown in Fig. 8-13. A disadvantage of the circuit with the single tuning capacitor is that the rotor of the tank capacitor is above ground potential and small changes in capacitance caused by bringing the hand near the tuning control causes detuning. This may be offset by substituting for the single capacitor a balanced split-stator capacitor with the rotor grounded as in Fig. 8-14. The equivalent bridge circuit is shown in Fig. 8-15. C_n is adjusted for a balance and is equal to C_{ga} provided C_1 is equal to C_2. The anode-cathode capacitance of the valve is across one half of the tuning capacitors. This causes unbalancing of the circuit and renders neutralization over a band of frequencies more difficult. A circuit which overcomes this drawback is shown in Fig. 8-16. The capacitor C_x is a small trimmer placed across one half of the tuning capacitor. The anode-cathode capacitance is in parallel with the other half of the tuning capacitor. Thus C_x can be adjusted to equal the anode-cathode capacitance and give neutralization. This anode-cathode capacitance however changes with frequency and will cause unbalancing. The range of neutralization is therefore limited. A capacitor C_x can also be used in the circuit of Fig. 8-12 to balance the

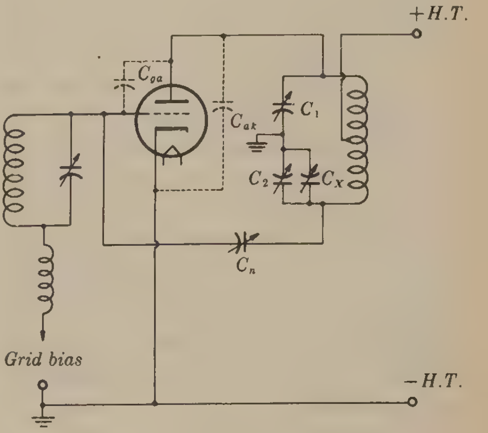


FIG. 8-16 Neutralized amplifier with balanced load circuit.

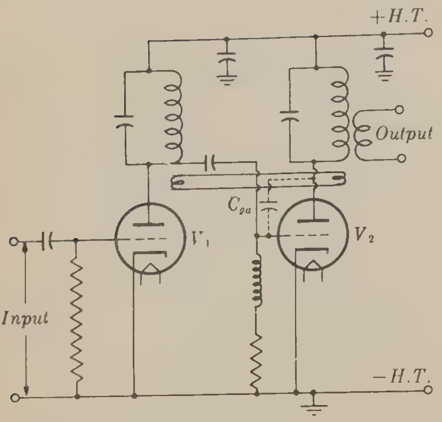


FIG.8-17 Neutralization by inductive feedback.

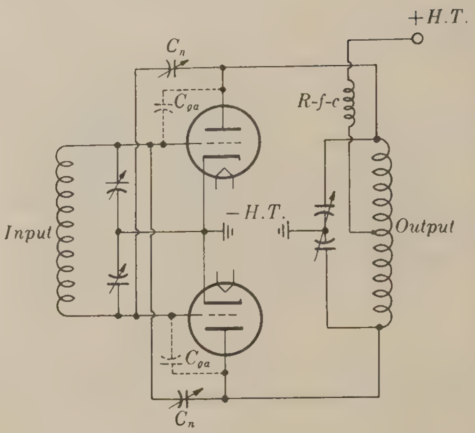


FIG.8-18 Neutralized push-pull amplifier.

anode-cathode capacitance in that circuit.

The balanced-bridge type of neutralization holds for a band of frequencies and is therefore usually to be preferred. However, inductive feedback can be used to counteract the feedback through the anode-grid capacitance. This can only hold for one frequency but is useful in reducing parasitic oscillations which may occur at some set frequency. The circuit in Fig. 8-17 illustrates this method of neutralization for V_2. There is link coupling between the output of V_2 and the input of V_2 via the output of V_1.

The energy fed back to the input of V_2 must be equal and opposite in phase to the energy fed back from the output tank of V_2 through C_{ga}. The mutual inductance is adjusted to neutralize the circuit for one particular frequency. If the frequency changes, the mutual inductance is changed so that a balance is no longer maintained. However, this type of neutralization is useful when the grid-anode capacitance of the valve is very small and a suitable balance cannot be obtained with circuits using a neutralizing capacitor.

In push-pull amplifiers two neutralizing capacitors must be used as shown in Fig. 8-18. The neutralizing capacitances are approximately equal to C_{ga}. With identical valves and a symmetrical circuit the neutralization is complete for all frequencies. Split-stator capacitors with the rotors grounded in the anode and grid circuits help to give good circuit balance. That is, the a-c voltages with respect to ground at corresponding points in both valve circuits have identical amplitudes. A push-pull circuit is useful at all frequencies but especially so at high frequencies where it can be satisfactorily neutralized.

In the push-pull circuit good neutralization can be obtained because the input capacitance, the output capacitance and the anode-cathode capacitance occur across both halves of the tuned circuit. Hence there is no detuning as the frequency of the input is changed.

The type of neutralizing capacitors used must be chosen to withstand the large a-c and d-c voltages which appear across the neutralizing capacitor. Thus the voltage rating of this capacitor must be very high. In a high-power transmitter this may be of the order of several thousand volts. The capacitance of the capacitor will depend upon the interelectrode capacitance of the valve or valves used, plus the stray capacitances introduced by the wires which connect the neutralizing capacitor into the circuit. The value chosen is roughly equal to these capacitances when the neutralizing capacitor is half meshed. The adjustment is usually made by means of a screw driver and, when correctly set, must be secured so that the neutralization is not changed by vibration.

The practical procedure for neutralizing an amplifier depends on the following principle. When the neutralizing capacitor is adjusted so that there is no feedback from the output circuit to the input, there can be no feed forward from the input to the output. This follows from the fact that the neutralizing circuit is a bridge circuit. When an a-c voltage is applied across the input to the bridge, no a-c, due to the output a-c, appears across the input. Hence, when an amplifier is neutralized for no feed forward, there can be no feedback.

To adjust the neutralization in a power amplifier stage, a direct-current meter is inserted in its grid circuit. The driving voltage is applied to the grid circuit with the H.T. removed and the filaments operating. Thus a reading will be obtained on the grid current meter since grid current flows. The magnitude of the grid current depends upon the amplitude of the driving voltage. The output tank capacitor is then tuned through its entire range and the meter observed. If the stage is not neutralized, a dip in the meter reading occurs as the tank circuit comes into resonance. Energy is being fed forward from the input to the output tuned circuit. At resonance the output circuit has a maximum circulating current. This forms a load which dissipates maximum energy and the input a-c is damped so that minimum grid current flows. A neon bulb or a small flashlight bulb loosely coupled to the anode tank circuit may also be used as an indicator. The oscillation of the anode circuit when it comes into resonance will cause the bulb to light up. If there is no dip in the grid current meter reading or the light bulb does not glow, the power amplifier is neutralized.

If neutralization is necessary the following procedure is employed. The anode circuit is tuned to resonance. This occurs at the minimum grid current point or the maximum brilliance of the lamp. The neutralizing capacitor is then adjusted by means of an insulated screwdriver until the grid current rises to its maximum value or the lamp goes out. Then the tank capacitor is tuned through its range to check the neutralization. In some cases, the feed forward will occur again at some other tuning position. If this happens complete neutralization over the range of the tuning capacitor may not be possible. However, the tank capacitor is again tuned for a dip in the meter reading and the neutralizing capacitor adjusted until the grid current comes up again. If complete neutralization is not possible, the neutralizing capacitor is tuned either for the optimum neutralization over the whole range of the anode tank capacitor or for complete neutralization at the operating frequency.

In a transmitter, a c-w signal is always used for the neutralization procedure to give a steady grid current. The neutralization is always adjusted on the highest frequency range of the transmitter. If it is neutralized on this range, it will seldom have to be re-neutralized on the lower frequency ranges.

In a push-pull stage, the two neutralizing capacitors must be tuned simultaneously in order to get neutralization. This method is usually used but its success depends upon the skill of the operator. If a more exact method is necessary each valve can be neutralized separately by first removing the other valve. In this way the two neutralizing capacitors are adjusted independently, according to the above procedure.

8.10 Parasitic oscillation. In an oscillator or amplifier the circuit conditions may be such that undesired oscillation takes place. This is called a *parasitic oscillation.* The energy required to maintain a parasitic oscillation is wasted so far as useful output is concerned. Hence an oscillator or an amplifier having parasitics operates at reduced efficiency. The presence of parasitics often causes the operation at the desired frequency to be erratic.

Parasitic oscillations can be divided into two classes according to the frequency. They occur at ultra-high frequencies and at frequencies below the circuit operating frequency. Parasitics are most likely to occur in circuits using large power valves because of the long leads, large interelectrode capacitances and relatively high values of transconductances.

The low-frequency parasitic oscillation is usually caused by the r-f chokes in the anode and the grid circuits resonating with the decoupling capacitors to produce a tuned-anode tuned-grid oscillator at a frequency of 500 Kc/s or less. The other tuned circuits have no effect on this since the reactances of the coils are small at the low frequency and so they act as short circuits. Thus the setting of the main capacitor has no effect on the frequency of the parasitic and shorting the anode tank circuit will not stop the oscillation. To eliminate this type of parasitic, the circuit should be arranged so that only one r-f choke is necessary either in the grid or the anode circuit. In push-pull stages a resistance in series with the r-f choke in the grid circuit damps out the oscillations. This resistor is called a *parasitic suppressor.*

The very-high-frequency parasitics are usually caused by the inductance of leads resonating with the stray capacitances in both grid and anode circuits. This again forms a tuned-anode tuned-grid oscillator. To remedy this a wave trap tuned to the parasitic frequency or a series resistor of about 50 ohms may be used to damp out the oscillation.

The parasitic resistor or parasitic suppressor is placed in series with the grid, screen or anode leads. The parasitic oscillatory circuits may be detuned by making the anode leads much longer than the grid leads so that they do not resonate at the same frequency.

8.11 Power supplies and controls. For a low-power transmitter or transceiver (12-25 watts output) a single power supply is usually used to operate all circuits. The H.T. voltage of 300-600 *V* is obtained from a high-vacuum twin-diode rectifier. For a 50-500 watt transmitter two power supplies are necessary. One supplies the low H.T. voltage of 300-600 *V* for the stages preceding the power amplifier and the other supplies the high H.T. voltage of 1000-2000 *V* for the power amplifier stage. The latter ordinarily consists of a high-voltage step-up transformer connected to two large mercury-vapour diodes which rectify the a-c input to give a high voltage output and deliver a large amount of power.

The power rating of the supply required to operate a transmitter depends not only on the r-f power which the transmitter is to produce but also on the over-all efficiency of the transmitter. In order to obtain a given amount of r-f power output, a very much larger amount of power must be provided by the power supply necessary. The final power amplifier stage is 50-75% efficient but the previous stages have a somewhat lower efficiency. The power required by the filament and control circuits must also be taken into account when calculating the power which the supply must provide. As an example, a low-frequency 500-watt transmitter with an a-c power supply may require 1800 watts in the valves and another 200 watts in the filament circuit. The total input is then 2000 watts for an r-f output of 500 watts. This transmitter is roughly 25% efficient in transforming the a-c power input to the supply into r-f power. The a-c power input may be supplied by a rotary converter operated from the ships d-c mains so that more power is supplied by the ships mains than is actually supplied to the transmitter because of the losses in the rotary converter. From the d-c ships mains, then, the efficiency is somewhat less than 25% in the production of the r-f power.

The *control circuits* of the transmitter are used to turn on and operate the set. A large number of controls are necessary some being on the front panel and some being inside the set. The tendency in more recent equipment is to minimize the number of outside controls to make the operation of the set less complicated. However the following discussion includes all the circuits and relays which act as controls.

A *main-line* switch, external to the set, is usually supplied for removing power entirely from the transmitter. This is normally closed and the transmitter is turned on by a switch on the panel. Two types of operation, *stand-by* and *normal*, must be provided for. In the stand-by position the transmitter is ready to be operated at a moment's notice. This means that the filaments are heated and either the H.T. is ready to be applied, or is already applied along with a paralyzing bias which keeps the valves cut off. The switching from stand-by to normal operation is done by a switch, key or press-to-talk switch on the microphone. A transceiver or transreceiver in stand-by position has the receiver operating. When the switch, key or press-to-talk switch is pressed the receiver is turned off and the transmitter turned on. In this case a single aerial is used for both transmitter and receiver. Thus great care must be taken to turn off and mute the receiver for its protection, before the transmitter is turned on. A special relay called a *break-in keying relay* is used for this purpose.

Several safety devices are incorporated in the control circuits. An *overload relay* protects the valves and power supply from being damaged by a short circuit or a breakdown in any part of the transmitter. This relay may be in the return path of the power supply so that all the power supply current flows through it or it may be in the power amplifier cathode circuit so that only the power amplifier current flows through it. In one case the whole transmitter and power supply are protected while in the other case only the power amplifier is protected. *Line fuses* are always used to protect the whole transmitter in case of failure in the overload relay or in case the overload relay does not protect the whole transmitter. If mercury-vapour valves are used in the power supply their filaments must be turned on before their a-c anode voltage is applied. To allow time for the filaments to heat up, a *time-delay relay* may be used to close a switch in the a-c line to the H.T. transformers.

The voltages used on most transmitters are quite high and therefore dangerous to maintenance personnel if not turned off during servicing. Consequently safety *door* and *panel* switches are used to automatically cut off the H.T. when the door or panels are open. These switches are spring loaded and are held closed by the pressure of the panel when it is properly in place.

8.12 A transmitter circuit. The transmitter (Fig. 8-19) is of medium power

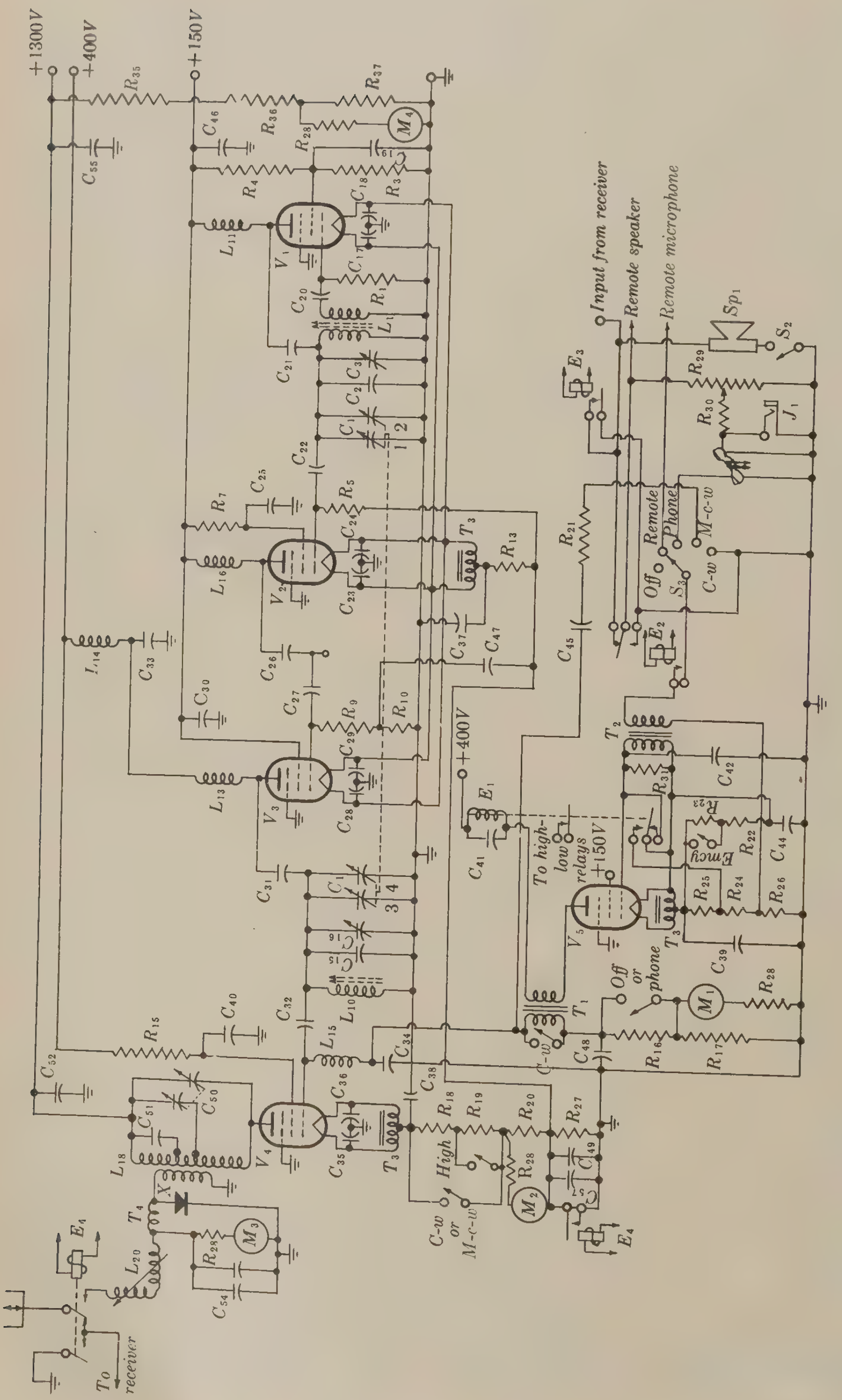


FIG. 8-19 A medium-power transmitter.

and is suitable for small patrol vessels. A receiver is used in conjunction with the transmitter. The transmitter frequency ranges are just below and above the broadcast band and there is both master-oscillator and crystal control. Transmission is on c-w with a power output of 100 watts, on m-c-w with 70 watts and on r-t with 30 watts.

The aerial is a quarter-wave Marconi grounded at one end. The aerial loading components permit selection and adjustment according to frequency and are housed in an aerial chassis. The same aerial is used for the receiver as well as the transmitter.

A. Master Oscillator

The oscillator is a parallel-feed tuned-anode oscillator employing a beam power tetrode V_1. L_1 is a variable iron-core inductance which provides feedback from the anode circuit to the grid circuit by inductive coupling. L_1 is adjusted for dial-tracking at the low-frequency end of the band. $C_{1\text{-}1}$ and $C_{1\text{-}2}$ are two parts of the main tuning capacitor in parallel with each other. C_3 is a trimmer for dial tracking at the high-frequency end of the band. C_2 is a negative-temperature-coefficient capacitor which compensates for changes in the capacitance of the tank circuit with changes in temperature. This improves the frequency stability of the oscillator. C_{17} C_{18} form an r-f by-pass for the filament. R_1 and C_{20} provide automatic grid bias. L_{11} and C_{21}, the r-f choke and blocking capacitor, are for the parallel-feed arrangement. R_3, R_4 form a voltage divider across the 150V supply to provide the correct screen voltage. C_{19} is a screen by-pass capacitor and C_{46} is a by-pass capacitor for the 150V supply. R_{13}, C_{37} provide cathode bias for V_1, V_2 and V_3. T_3 is a filament transformer for the filaments of V_1, V_2, V_3, V_4 and V_5. The three secondary windings of T_3 are shown, the primary being supplied with 115V 60 c/s a-c.

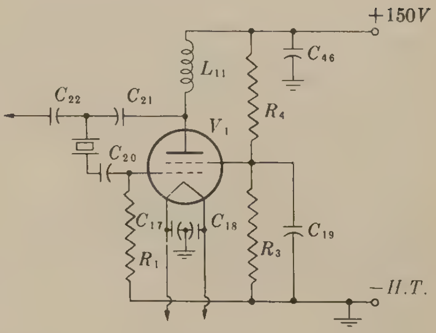


FIG. 8-20 First stage operated as a crystal oscillator.

B. Crystal Oscillator

The crystal oscillator is a parallel-feed Pierce oscillator employing the same beam power tetrode V_1, the circuit being changed to that shown in Fig. 8-20 by a switching arrangement. L_1 and associated capacitors are switched out.

C. Multiplier

This stage employs a beam power tetrode V_2. On lower frequency bands the circuit is a choke-capacitance-coupled untuned power amplifier. R_5 and C_{22} are the grid-leak resistor and coupling capacitor. L_{16} is the anode load inductor. C_{23} and C_{24} form an r-f by-pass for the filament. R_7 is a screen voltage dropping resistor and C_{25} is a screen by-pass capacitor. On high-frequency bands the circuit is changed to become a parallel-feed tuned-anode multiplier (Fig. 8-21). A tuned circuit consisting of a variable inductance L_5, a trimmer capacitor C_9 and a part of the main tuning capacitor $C_{1\text{-}5}$, all in parallel, is switched in. The inductance L_5 is tuned on the low-frequency end of the band and the trimmer C_9 is tuned on the high-frequency end for dial tracking. When the tuned anode circuit is employed the multiplier is used as a tripler. C_{26} is the

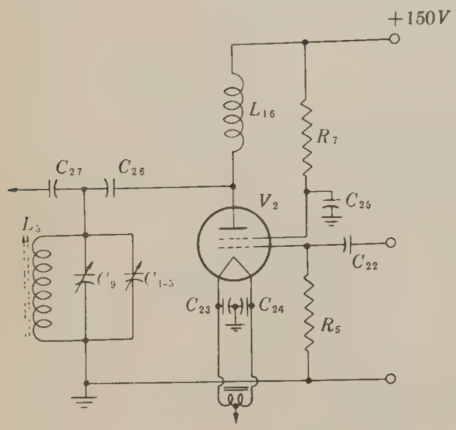


FIG.8-21 Second stage operated as a frequency multiplier.

d-c blocking capacitor for parallel feed.

D. Driver

V_3 is a beam power tetrode. The circuit is a parallel-feed tuned-anode r-f power amplifier. R_9 and R_{10} provide grid-leak bias along with C_{26}, C_{27}. C_{47}, a large by-pass capacitor, maintains the bias steady across R_{10}. Therefore R_9 only is an a-c load on the previous stage. The tuned anode circuit consists of L_{10}, C_{16}, C_{15}, $C_{1\text{-}3}$ and $C_{1\text{-}4}$ in parallel. L_{10} is a variable iron-core inductance for dial tracking at the low-frequency end. $C_{1\text{-}3}$ and $C_{1\text{-}4}$ are sections of the main tuning capacitor. C_{15} is a basic capacitor and C_{16} a trimmer for dial tracking at the high-frequency end. The choke L_{13} and the blocking capacitor C_{31} form the parallel-feed arrangement. C_{30} is the screen by-pass capacitor. C_{28} and C_{29} provide r-f by-pass for the filament. L_{14} and C_{33} form an r-f filter to keep r-f out of the 400V supply.

E. Power Amplifier

This final stage employs a beam-power tetrode V_4 which has higher power-handling capabilities than the preceding valve. The circuit is a Class C r-f power amplifier with a series-feed tuned anode output circuit. There is grid modulation of the amplifier via T_1. L_{18} is the inductance of the tank circuit. C_{50} is a main tuning capacitor for the tank circuit and is controlled from the front panel. One or two sections of the capacitor are used according to the frequency. C_{51} is a basic capacitance used on some frequencies. L_{15} is an r-f choke to provide a high impedance from grid to ground. C_{34} is an r-f by-pass for the output of the modulation transformer T_1. R_{16} is a grid-leak bias resistor used on c-w and m-c-w. The switch OFF-PHONE shorts out R_{16} for r-t. R_{17} is a small meter shunt. R_{28} is a series meter resistor. M_1 is a multimeter which reads power amplifier grid current when switched in position 1. C_{48} is an a-f by-pass for the grid resistor R_{16} and the meter. C_{38} is an r-f filament by-pass. R_{18} is a cathode bias resistor used for r-t only. The C-w or M-c-w switch shorts R_{18} and R_{19} out for c-w and m-c-w. R_{19} is a cathode bias resistor which is switched in by opening the switch across it for r-t transmission if a lower power than 30 watts is required. The low power is provided by lowering the high H.T. voltage by means of a switch controlling the number of turns to be used on the H.T. transformer primary. R_{20} is a small meter shunt so that M_2, the multimeter when switched in position 2, reads the power-amplifier cathode current. M_4 is the multimeter switched in position 4 to read H.T. voltage. R_{35}, R_{36}, R_{37} form a voltage divider across the H.T. supply, their purpose being to enable the meter to serve as a high-voltage voltmeter. C_{52} and C_{55} are H.T. by-pass capacitors. C_{35} and C_{36} are r-f filament by-pass capacitors.

F. Aerial Circuit

The output from the power amplifier is transformer coupled to the aerial. L_{20} is a variometer in series with the output coupling coil and the aerial is tuned to bring the aerial into resonance. The multimeter M_3 when switched into position 3 provides an aerial current meter. Transformer T_4 couples r-f into the meter circuit. A copper-oxide rectifier X produces a d-c output for the meter so that the meter reading is proportional to the r-f current flowing in the aerial circuit. C_{54} is a two-section capacitor which forms an r-f filter for the output of the rectifier for the aerial current meter. E_4 is a break-in keying relay. The two contacts shown to the left of the relay coil E_4 in the aerial circuit are in the receiver position. The aerial is connected to the receiver input. A third contact across R_{27} is open so that paralyzing bias is applied to the transmitter. This paralyzing bias is discussed in subsection G. When the relay coil E_4 is energized during transmission by closing the key or press-to-talk switch, the second contact shown to the left of E_4 in the aerial circuit closes and grounds the receiver input terminals. The first contact to the left of E_4 closes to connect the aerial to the transmitter and the third contact closes to short out R_{27} and remove the paralyzing bias.

G. Keying Circuit

The key or the press-to-talk switch on the microphone operates the keying relay E_4 as well as E_3 and E_2. When the key is open the cathode circuits of the master oscillator V_1, multiplier V_2, driver V_3 and power amplifier V_4 are returned to ground through R_{27}. The high voltage on the power amplifier causes sufficient cathode current to produce a large voltage across R_{27}. This provides a paralyzing bias which cuts off the master oscillator and driver. Since the grid leak R_5 of the multiplier is connected to the upper end of R_{27}, the paralyzing bias does not affect the grid-to-cathode voltage of the multiplier. It does, however, reduce the anode voltage on the multiplier. When the key is pressed R_{27} is shorted due to the energizing of E_4 and normal bias is applied to all stages so that they are operating.

H. The Modulator

The modulator is an a-f power amplifier transformer-coupled by T_1 to the grid of the r-f power amplifier for grid modulation when operating on r-t. For m-c-w the circuit operates as a tuned-grid a-f oscillator. For c-w the modulator input and output are shorted. The type of operation is determined by the position of the switch S_3. The circuit components R_{22}, R_{23}, R_{24}, R_{26} and C_{44} in the cathode circuit are discussed in the time-delay subsection *I*. R_{25} and C_{39} provide cathode bias.

1. M-C-W OPERATION.

For m-c-w operation the switch S_3 is in the M-c-w position. The secondary of the modulation transformer T_1 is connected back to the input side of the microphone transformer T_2 through the feedback line containing R_{21} and C_{45}. R_{21} and C_{45} are chosen to provide the correct quantity and phase of feedback to cause audio oscillation. This is a tuned-grid oscillator since T_2 controls the circuit.

2. R-T OPERATION.

For r-t operation S_3 is switched to the PHONE position so that the microphone is connected to the microphone transformer T_2 primary. R_{31} is shunted across the secondary of T_2 to give a more uniform response to voice frequencies. C_{42} and C_{44} in series also shunt the T_2 secondary to give more uniform response. E_2, the microphone relay, is energized when the press-to-talk switch is closed on the microphone handset so that the one set of E_2 contacts shown below the coil E_2 close to apply the microphone output to the input transformer T_2. The second set of E_2 contacts shown above the coil

E_2 close in the downward position. As a result the input from the receiver is grounded via these E_2 contacts and also via the E_3 contacts which are closed at this time, since E_3 is also energized when the press-to-talk switch is closed.

When the press-to-talk switch or the key is not closed the E_2 and E_3 contacts are in the position shown. The input from the receiver is connected via the E_2 contacts to the earphone component of the handset. The input from the receiver is also connected at all times to the speaker Sp_1 which may be turned on or off by means of the switch S_2.

R_{26} provides a polarizing voltage for the microphone circuit and provides some cathode bias for V_5 before the time-delay relay operates. J_1 is an earphone jack for an extra set of earphones. R_{30} is a series resistor to reduce the input from the receiver to the earphones. R_{29} is a potentiometer to act as a volume control for the receiver input to the earphones.

The REMOTE position of S_3 permits the microphone of a remote handset to feed into the modulator and permits the input from the receiver to be connected to a remote speaker. In this case the press-to-talk switch on the remote set controls the E_1, E_2, E_3 and E_4 relays.

3. C-W OPERATION.

On c-w operation S_3 is switched to the C-w position. When the key is pressed E_2 and E_3 are energized. The receiver is grounded as before. At the same time the primary of the microphone transformer T_2 is grounded through S_3. Thus the modulator receives no input during c-w transmission. The modulator output is also shorted by the c-w switch across the secondary of T_1.

I. The Time-Delay Relay

The time-delay relay is provided to allow the filaments of the mercury-vapour valves in the H.T. power supply to heat up for about 15 seconds before the H.T. transformer is energized. The time delay must operate also before V_5 can act as a modulator. Before the transmitter is turned on, the coil of the relay E_1 which is in the anode circuit of V_5 is not energized and the contacts for E_1 are in the position shown in the diagram. All filaments and the 400V d-c power supply are turned on when S_3 is switched from the off position to any of its other three positions. Current now flows through V_5 and produces a voltage drop across R_{25}, R_{24} and R_{26} which raises the cathode above ground potential. The grid is connected to the top end of C_{44} which is not charged initially. This causes the valve to operate close to cut-off so that very little current is flowing through the coil of E_1. As C_{44} charges through R_{22} and R_{23} the potential of the control grid gradually rises until sufficient current flows through V_5 to trip E_1 which requires about 20mA d-c. Now the contact of E_1 which is connected to the bottom of R_{22} changes position so that the bottom of R_{22} is connected to the bottom of R_{25}. Thus, R_{22} and R_{23} in series are placed in parallel with R_{25} to provide normal operating bias for V_5. The voltage across R_{24} and R_{26} no longer provides bias but effectively lowers the anode potential slightly. The remaining contacts of E_1 close simultaneously to

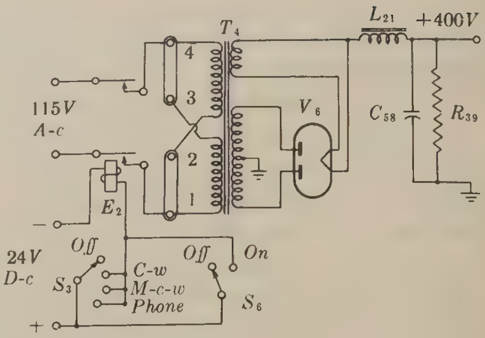


FIG.8-22 Low H.T. supply.

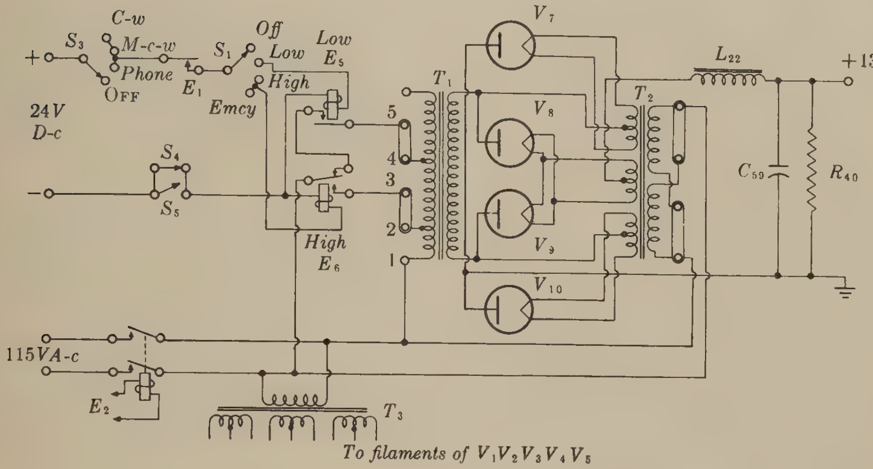


FIG. 8-23 High H.T. supply.

permit the high H.T. voltage to be applied to the power amplifier.

Emergency operation is controlled by the switch EMCY. across R_{23}. When the switch is closed R_{23} is shorted out thereby reducing the time constant in the delay circuit to approximately one-quarter of its normal value. The transmitter can be brought into operation in about 4 seconds in this case.

J. Power Supplies

The transmitter is designed to operate from 24V d-c and from 115V 60 c/s a-c or 230V 60 c/s a-c. If only 24V d-c is available initially, the 115V a-c supply may be obtained by using a rotary converter. If only 115V a-c is available initially the 24V d-c supply may be obtained by using a suitable rectifier circuit.

The 24V d-c supply is used to energize the bobbins of the relays E_1, E_2, E_3 and E_4. The 115V a-c is used to energize the H.T. and filament transformers.

1. 400V H.T. Power Supply.

The power supply is shown in Fig. 8-22. When switch S_3 is turned to C-w, M-c-w or Phone position the 24V d-c supply energizes E_2 which in turn completes the 115V a-c circuit to T_4. The switch S_6 in a remote control unit performs the same function when turned to the On position.

The power supply is a full-wave rectifier using a double diode V_6. The filter circuit consists of L_{21} and C_{58}. R_{39} is a bleeder resistor.

If the a-c input is 230V the two primary windings of T_4, which are in parallel for the 115V input, are put in series by removing the links 1, 2 and 3, 4 and connecting a link between 2 and 3.

2. 1300V H.T. Power Supply.

The power supply is shown in Fig. 8-23. When the switch S_3 is turned to C-w, M-c-w or Phone position E_2 is energized as shown in Fig. 8-22. The 115V a-c is then applied to the filament transformer T_3, which supplies the transmitter valve filaments, and to the filament transformer T_2 which supplies the filaments of the rectifier valves V_7, V_8, V_9 and V_{10}. The 115V a-c is not applied to the H.T. transformer T_1 until the following switching operations are carried out.

(i) S_3 must be turned to C-w, M-c-w or Phone.

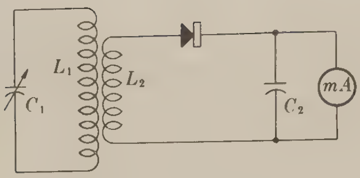


FIG.8-24 Absorption wavemeter employing crystal detector.

(ii) Either the *gate switch* S_4 or the *lock switch* S_5 must be closed. The gate switch S_4 is a spring-loaded switch which automatically opens when the front panel is removed. If it is necessary to turn on the power supply when the front panel is off, as in servicing, the lock switch must be pulled forward to close it. When the front panel is replaced the panel closes the gate switch S_4 and pushes in the lock switch S_5, thus opening it.

(iii) The contacts of the time-delay relay E_1 must be closed.

(iv) The switch S_1 must be turned to LOW, HIGH or EMCY. In the LOW position the 24 V d-c energizes E_5. The relay contacts move to cause the 115 V a-c to be applied between terminals 4 and 1 on the primary of T_1. The output d-c of the power supply is only 650 V. This LOW position is used for preliminary tuning of the transmitter. In the HIGH position the relay E_6 is energized. Its contacts move to cause the 115 V a-c to be applied between terminals 2 and 1 on the primary of $T1$. The output d-c is now 1300 V.

The H.T. power supply employs four mercury-vapour diode rectifier valves in a bridge arrangement. L_{22}, C_{59} form a choke-input filter. R_{40} is a bleeder resistor to stabilize the voltage and discharge C_{59} when the unit is turned off.

8.13 Wavemeters. To determine the wavelength or the frequency of the radiation from a given circuit, an instrument known as a *wavemeter* or *frequency meter* is employed. The same meter may be calibrated to read either frequency or wavelength since, if one is determined, the other is calculated from the relationship:

wavelength × frequency = velocity

Thus $f \times \lambda = v$

where f = frequency in c/s
λ = wavelength in metres
v = velocity of propagation
$= 3 \times 10^8$ meters/sec.

$$\lambda = \frac{v}{f}$$

$$= \frac{3 \times 10^8}{f} \text{ metres.}$$

A. Absorption Wavemeters

A simple absorption wavemeter circuit is shown in Fig. 8-24. It consists of a resonant circuit L_1C_1 which may be tuned by C_1 to resonate at the frequency to be measured. A second coil L_2, coupled to L_1, is connected in series with a crystal detector and a d-c milliammeter. C_2 is an r-f by-pass capacitor in parallel with the meter. A pointer connected to the rotor of C_1 moves over a calibrated scale on the front of the meter.

The wavemeter is placed close to the transmitting circuit, the frequency of which is to be measured. Some of the energy radiated from the transmitter is absorbed by the circuit L_1C_1. When this circuit is tuned to resonance at the transmitter frequency, the circulating current in L_1C_1 becomes a maximum. This in turn induces a maximum e.m.f. in L_2. The r-f voltage across L_2 is rectified by the crystal and the d-c component of the rectified current flows through the meter. C_1 therefore is adjusted for maximum current through the meter to indicate resonance in L_1C_1. The frequency or wavelength of the radiation is then read from the calibrated scale on the meter panel.

A practical absorption wavemeter

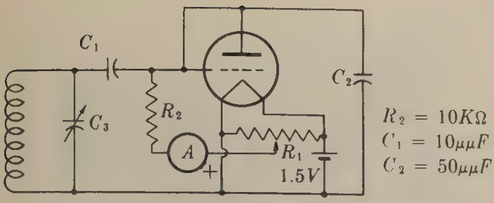


FIG.8-25 Wavemeter employing vacuum-tube voltmeter.

which has a wide frequency range is shown in the circuit diagram of Fig. 8-25 and in the photograph of Fig. 8-26. It contains a vacuum-tube voltmeter in place of the crystal shown in the simpler wavemeter described above. To cover the entire frequency range desired, various coils are used to resonate with the one capacitor, one coil for each band of frequencies. When the wavemeter has more than one coil the capacitor scale is not calibrated to read frequency but the scale is numbered. The frequency corresponding to any scale reading is obtained from a frequency calibration chart. There is one frequency chart for each coil.

The vacuum-tube voltmeter uses a triode valve connected as a diode. There is a 1.5V battery to heat the directly-heated cathode. A potentiometer R_1 is connected across the filament terminals. The d-c galvanometer is connected to the variable arm on the potentiometer and in this way gives a zero adjustment for the meter with no signal input. For any particular setting of R_1 part of the cathode, is negative with respect to the anode and some current will flow through the valve and meter. The zero of the scale on the galvanometer is displaced from the true zero so that with a small amount of current the meter reads zero. Thus R_1 can be varied to produce the zero reading on the galvanometer for no signal input.

The wavemeter coil is inductively coupled to the circuit whose a-c frequency is to be measured. The voltage set up across C_3 is then applied through C_1 to the diode which is in parallel with the load resistor R_2 and the meter. On the positive half cycles the diode conducts through R_2 and the galvanometer to produce a meter reading and to charge C_1 to almost the peak value of the a-c voltage across C_3. On the negative half of the cycle C_1 begins to discharge through R_2 and the meter, thus maintaining the meter reading. If the time constant C_1R_2 is long compared with the period of the a-c applied, then the capacitor discharges slightly and on the next positive half cycle will charge again. Thus the current which flows through the valve is just enough to make up for the discharge of the capacitor through R_2 and the meter during the

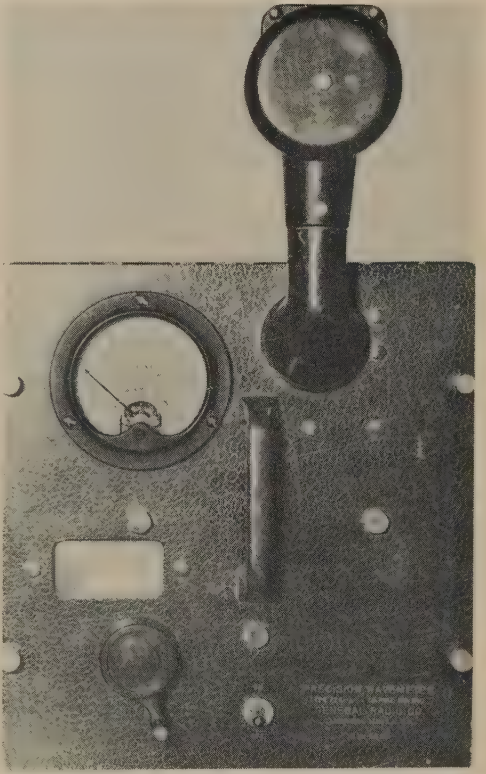


FIG.8-26 Front panel of vacuum-tube wavemeter.

non-conducting half cycle and the meter reading is proportional to the amplitude of the voltage across C_3.

As the capacitor in the wavemeter is tuned to resonance, the amount of energy absorbed in the wavemeter input circuit is a maximum and the meter reading is a maximum. Thus the setting of the capacitor, when it is tuned for a maximum meter reading gives a reading from which the value of the frequency being measured is obtained. The coupling between the circuit and the wavemeter coil can be varied so that the meter reading at the maximum is not too large. The amount of energy taken out of the circuit may affect the frequency and so the smallest amount of coupling that will give a useful indication of the wavemeter is desirable. While tuning for a final reading, the position of the wavemeter relative to the circuit must remain fixed.

The use of the capacitor C_2 across the diode is explained as follows. Without C_2 there are two capacitances in series across the tuned circuit. These are C_1 and the anode-cathode capacitance C_{ak} of the valve. C_1 and C_{ak} are of the same order of magnitude. The effective capacitance of C_1 and C_{ak} in series depends to a large extent upon the valve interelectrode capacitances. As the valve ages or a new valve is used, this effective capacitance is changed. Since this effective capacitance is in parallel with C_3, the frequency of resonance of the wavemeter for a given setting of C_3 is changed. However, the calibration of C_3 should be the same for all valves used. C_2 is placed in the circuit to make the tuning independent of the interelectrode capacitance. Since C_2 is $50\mu\mu F$ and is large compared with C_{ak} a change of a few $\mu\mu F$ in parallel with it will have no effect in the circuit and the calibration of C_3 remains reasonably constant for all valves.

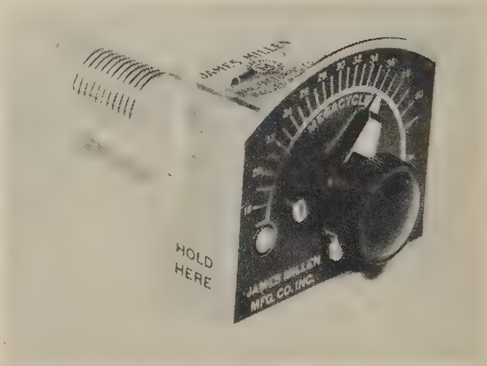


Fig.8-27 Reaction type of absorption wavemeter.

The actual frequency reading is determined indirectly through a calibration chart. An accurate micrometer gauge rotates the capacitor C_3. Thus a very fine adjustment is possible and an accurate reading of the capacitor setting is obtained. The calibration chart consists of a graph of frequency in Kc/s against capacitor setting. From this chart the frequency of an a-c can be determined to within 0.25%. To cover a large frequency range several coils of different inductances are used. Each coil has its own calibration chart.

One simple type of absorption wavemeter has no indicating meter. It consists of a coil and a variable capacitor as shown in Fig. 8-27. A resonance indication is obtained by observing the anode current meter in the transmitter stage whose frequency is being checked. In this case, when the wavemeter is tuned to resonance, the increased loading on the tank circuit of the transmitter stage causes a rise in the anode current of the stage. This effect is a maximum when the wavemeter is exactly tuned to the same frequency as that of the transmitter stage. This method of indication is called the *reaction* method. To cover a wide range of frequencies several coil and capacitor units are provided as shown in Fig. 8-28.

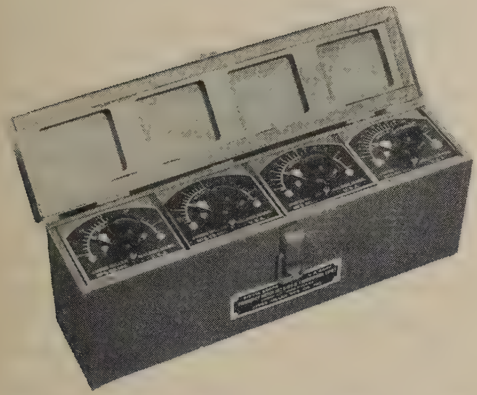


FIG.8-28 Wide range reaction type wavemeter.

B. The Heterodyne Wavemeter

This type of wavemeter generates oscillations of a known frequency which are combined with the oscillations of the circuit being tested and the beat frequency between the two is detected. This is called a *heterodyne wavemeter.* The circuit diagram is shown in Fig. 8-29.

The wavemeter is a combined oscillator and detector. A single battery is used for H.T. and L.T. supplies. The L.T. is varied by a rheostat to provide the correct filament voltage. The signal from the circuit being tested is introduced into the input circuit through the coupling coil. This signal is combined in the tuned circuit with the oscillation produced by the wavemeter and the resultant beat detected by means of grid detection in the triode. A current of frequency equal to the difference between the signal and the oscillation frequencies flows through the headphones. If this difference frequency is in the a-f range, an audio note is heard. When the variable capacitor is set so that there is zero beat, that is, when the frequency of the note has decreased to zero, then the frequency reading on the wavemeter is also the frequency of the circuit being tested. This is a very sensitive method and gives very accurate measurement of frequency.

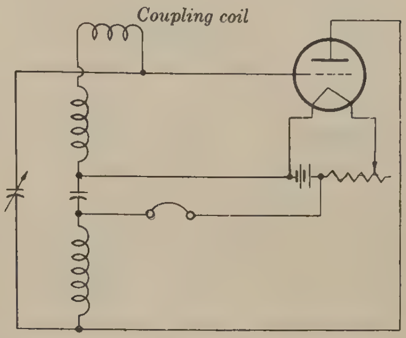


FIG.8-29 Heterodyne wavemeter circuit.

8.14 Tuning the transmitter. Before tuning the actual transmitter circuits, the following preliminary adjustments are made.

(i) The transmitter is switched to C-w. Tuning is always done on c-w to give steady conditions. If only r-t is available for the transmitter the microphone is disconnected or muffled.

(ii) The filaments are switched on. Sufficient time must be allowed to elapse to permit the filaments to heat up.

(iii) A reduced H.T. is switched on to the high-power stages of the transmitter. This is necessary in a high-power transmitter stage so that high-voltage H.T. is not applied to an untuned stage; otherwise an excessive anode current might flow to burn out the valves. The switch which provides the reduced H.T. is usually labelled the HIGH-LOW, OPERATE-TUNE or OPERATE-ADJUST switch. Sometimes internal knife switches are provided to break the H.T. to the high-power stage. These switches are closed after all previous stages are tuned so that H.T. is only applied for a short time before the final circuit is tuned.

(iv) The aerial is grounded. This is done in some cases so that no radiation occurs at an undesired frequency. It also prevents the aerial from having any effect on the circuit being tuned.

After the preliminary adjustments have been made the transmitter is tuned stage by stage. In many transmitters an approximate tuning is made by the adjustment of inductances. Movable taps are provided which are set by hand. Consequently before turning on the H.T., the taps are set on the coils at the approximate value of inductance recommended by the manufacturer's pamphlet. The pamphlet gives only approximate values but these can be used to advantage for rough settings. Since the H.T. voltages used are high enough to cause serious injury, care must be taken to turn off the H.T. before adjusting any coil taps by hand. The fine tuning is done with the variable capacitor. After the fine tuning, the capacitor setting is usually fixed by a dial lock. Some transmitters have a wave-change mechanism which enables several fixed settings of the one capacitor to be made. Several stops are employed on the one capacitor, so that the transmitter can be turned to any one of several preset frequencies. In either case, all the locks and settings are loosened when fine tuning is carried out.

The tuning of the transmitter starts with the master oscillator or crystal oscillator. If it is a master oscillator, a wavemeter must be used to check the frequency. The wavemeter is set to the desired frequency. The wavemeter coil is coupled to the tank circuit of the oscillator and the oscillator tuning capacitor tuned for a maximum wavemeter reading. In doing this the wavemeter should be coupled as loosely as possible so that the frequency of the oscillator will not change when the wavemeter is removed. Loose coupling also ensures that the wavemeter will not be overloaded when the transmitter is turned on.

If a crystal oscillator is used with a tuned circuit in the anode, this circuit must also be tuned. No wavemeter is necessary since the crystal sets the frequency. However no oscillations occur unless the tuned circuit is tuned near the crystal frequency. Proper tuning is indicated by a dip in anode current. Hence a cathode or anode current meter serves as a tuning indicator. The setting at the exact minimum of the current dip is not the final adjustment. At this setting the crystal oscillator is very unstable and loading caused by tuning the following stages may damp the oscillations in the tank circuit. This decreases the amount of feedback and may stop oscillation. If the set is turned off and on again at this setting of the tank circuit, the feedback may not be enough to start oscillations. Consequently to ensure starting of the oscillator and to ensure that loading caused by tuning the following stages does not stop oscillation, the circuit is detuned slightly from the resonant frequency. The detuning may be on the high-frequency or low-frequency side of resonance depending on the type of oscillator circuit.

The amplifiers or multipliers following the oscillator are now tuned to the incoming signal from the oscillator. If a current meter in the anode or cathode of the stage is used, a dip in meter reading indicates the anode circuit is tuned to resonance with the incoming signal. This is true of a multiplier as well as an amplifier. However, care must be taken with a multiplier to tune to the harmonic desired because a dip will result in tuning to any harmonic. In most cases, there is no ambiguity because the range of the tuning circuit includes only the desired harmonic frequency. Therefore the one minimum obtained is the correct one. Another indication of correct tuning which is perhaps more accurate may be used in an amplifier that is followed by another Class *C* stage with a grid or cathode current meter present. In this case when the amplifier is tuned, a maximum a-c voltage is applied to the grid of the next valve and causes the grid current in the next valve to reach a maximum. This maximum can be observed on a grid or cathode current meter.

By either of these methods the transmitter is tuned from stage to stage up to the aerial circuit. If there is a neutralization adjustment, the r-f circuits up to the neutralized stage are tuned and neutralization adjustments made as described in section 8.9. After the final power amplifier and aerial circuits are given a preliminary tuning with the reduced H.T. voltage on the power amplifier, final tuning is carried out with the full H.T. on.

8.15 Aerial coupling and tuning. The transmitting aerials used on low-frequency transmitters are usually short in comparison with the wavelength of the transmission. To make the aerial resonant, a certain amount of variable inductive loading is necessary. The tuning of this variable inductance determines the resonant frequency of the aerial circuit which is tuned to the transmitter frequency. Tuning the aerial brings the loading inductance and the aerial inductance into resonance with the aerial capacitance and stray capacitance to ground.

The aerial is coupled to the r-f power amplifier. Four different methods of coupling are shown in Fig. 8-30. Each of these methods has incorporated in it one or more variable components which are designed to make the aerial tune over the transmitter range of frequencies.

In Fig. 8-30 (a) the aerial is part of the tuned circuit of the power amplifier so that the problem of tuning is a simple one. When the circuit resonates the current circulating in the aerial circuit is a maximum. This current flows in the aerial and the maximum of energy is radiated. In tuning the circuit the drive is applied to the grid circuit and the capacitor C_2 set at half capacitance. The rough tuning is then done by rotating S_1 until a dip in the cathode current meter is observed. The position of minimum current is selected and an exact minimum is then obtained by fine tuning of C_2. The whole circuit is designed so that when the aerial is tuned it provides the correct load and the valve operates at its maximum efficiency.

For an indication of resonance in a separate aerial circuit, the tuning meter in the cathode or anode circuit of the final power-amplifier stage is usually used. The transmitter has been previously tuned so that the power amplifier is applying the required r-f to the aerial. When the aerial comes into resonance it acts as an open oscillator to radiate energy in the form of electromagnetic waves. If the r-f amplifier is operated Class *C* the extra power drawn from the circuit causes the d-c current through the valve to increase. A maximum cathode or anode current in the r-f amplifier shows that the aerial is in tune.

In the Fig. 8-30 (b), the aerial is attached directly to the coil in the power amplifier tank circuit. The impedance of the aerial is very low. For example, the impedance of a quarter-wave Marconi aerial is about 35 ohms. Since the impedance of the tank circuit is much more than this, the aerial tap is placed at a point close to ground where the impedance to ground matches the impedance of the aerial. The coil then acts as an autotransformer to

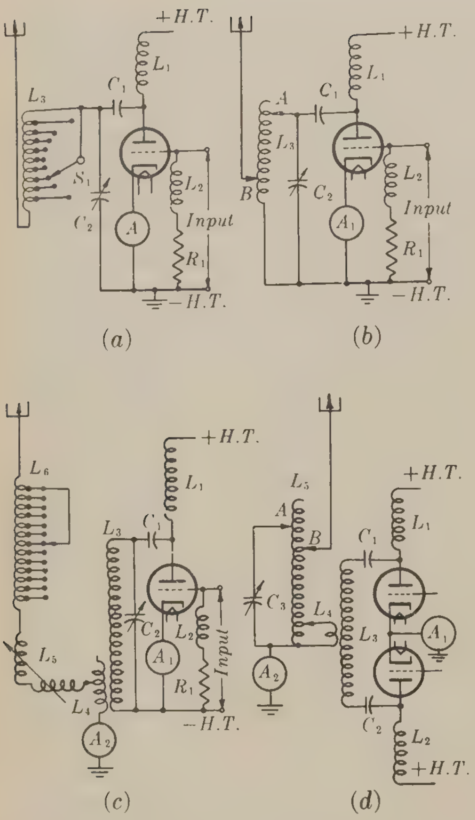


FIG.8-30 Methods of coupling aerial to power amplifier (a) Aerial part of tuned anode circuit (b) Direct coupling to p-a tank circuit (c) Mutual inductive coupling (d) Link coupling to aerial tank circuit.

match the aerial to the power amplifier tank circuit for maximum transfer of energy. In tuning this circuit, a milliammeter in the cathode circuit of the power amplifier is used. With the input to the power amplifier applied, the tap *B* removed and the capacitor C_2 at half capacitance, the inductance is varied by changing the tap *A* until an approximate minimum meter reading is obtained. Then C_2 is finely tuned to an exact minimum. Now the tap *B* is placed near the lower end of L_3. The aerial loads and also detunes the power amplifier tank causing a rise in cathode current. The capacitor C_2 is again tuned for a minimum meter reading to re-establish the correct tuning. This time the value of the minimum current, which is higher due to loading, is noted. The tap *B* is then moved up or down, each time tuning C_2 again to a minimum meter reading. The setting of the tap *B* which gives the largest value of minimum meter reading is the correct one. At this position the maximum transfer of energy to the aerial occurs.

When the aerial circuit is separate as in Fig. 8-30 (c), mutual inductive coupling is used to excite the aerial. The inductances L_3 and L_4 form the primary and secondary of a transformer which is used to match the aerial to the power amplifier tank circuit. The variable tap on L_4 varies the turns ratio to obtain this match. In order to make the aerial resonant, a loading inductance made up of L_6, L_5 and part of L_4 is used. The main inductance L_6 is the coarse adjustment and the variometer L_5 is the fine adjustment for the aerial circuit. The procedure in tuning is as follows. The tap on L_4 is placed near the grounded end to give a small amount of coupling to the aerial. This is to ensure that the power amplifier is not overloaded by too close coupling as L_6 is adjusted to tune the aerial. The position of the tap on L_6 is adjusted for a maximum reading of the power amplifier cathode current meter A_1. The aerial current meter A_2 will also read a maximum but since only a small amount of coupling is used the reading may be too small to be noticed. Now, since the aerial loads and detunes the power amplifier tank, the capacitor C_2 is returned for a dip in cathode current. The tap on L_4 is moved to increase the coupling until the rated current for full load on the power amplifier valve is obtained. Each time L_4 is adjusted, L_5 is tuned for a maximum cathode current and C_2 is retuned for a dip in cathode current. A check on the tuning is performed by moving

L_5 slightly in either direction to see if a greater aerial current flows through the aerial current meter. Movements of L_5 to right and to left should decrease the aerial current and L_5 is then returned to the maximum reading. However a slight adjustment of L_5 may increase the aerial current. If this is true then L_5 is tuned to the maximum.

In Fig. 8-30 (d) another type of matching circuit is illustrated. A fixed coupling coil L_4 is used to couple the power amplifier to the aerial circuit. The r-f is fed to a parallel tuned circuit made up of C_3 and the portion of L_5 from A to ground. The position of C, the coupling coil tap, and the position of A, the tuned-circuit inductance tap, determine the turns ratio of the autotransformer L_5. The correct adjustment of these taps matches the tuned circuit to the coupling coil L_4 for the maximum transfer of energy. The tap A is adjusted so that it gives the correct amount of inductance to tune within the tuning range of C_3. The aerial is then tapped on to a point B on L_5 so that the impedance from B to ground matches the aerial impedance. The matching autotransformer here has its primary from A to ground and the secondary from B to ground. The aerial current meter indicates when the aerial is matched to the tuned circuit. Resonance in the aerial tuned circuit is indicated by the power amplifier cathode current meter as a maximum reading.

The adjustment of this circuit is difficult. Each of the three taps on L_5 changes the effective circuit values in the tuned circuit. However, the general procedure in tuning the aerial is as follows. The coupling tap C is placed at a minimum coupling point. The parallel circuit can then be tuned to resonance as indicated by a rise in cathode current in the power amplifier. The tap A is set to a position which allows C_3 to tune the parallel circuit at a small capacitance point. The small amount of capacitance at resonance is desirable so that the a-c voltage developed will not be enough to cause arcing between the plates of the capacitor. Then the tap B can be varied until it is at a point where maximum aerial current is obtained. The capacitor C_3 is tuned with every adjustment until a maximum in power amplifier cathode current is observed. The coupling is then increased by means of tap C until the rated loading of the power amplifier is reached.

8.16 Microphones. A microphone is a device which is used to convert sound vibrations into electrical oscillations. The compressions and rarefactions which comprise the sound waves are made to fall upon a flexible diaphragm. These variations in air pressure cause the diaphragm to vibrate at the frequency of the vibrations in the sound wave. The movement of the diaphragm is then made to generate or to vary an electric current at the same audio frequency. The a-f current produced by the microphone is usually of small amplitude and therefore must be amplified. Hence the output of the microphone and its immediate circuit is generally applied to the grid circuit of a valve amplifier.

Five types of microphones which are in common use are described in the following paragraphs.

A. Carbon Microphone

The carbon microphone shown in Fig. 8-31 consists of a cylindrical metal cup filled with loosely packed *carbon granules.* The open face of the cup is fitted with a flexible metal diaphragm which is in contact with the carbon granules but insulated from the metal cup. This structure is called the *carbon button.* The microphone shown is called a *single-button* carbon microphone. There are two electrical leads from the microphone, one

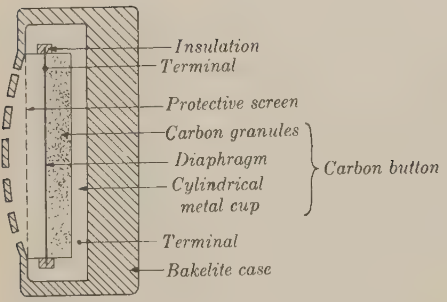


Fig. 8-31 Single-button carbon microphone.

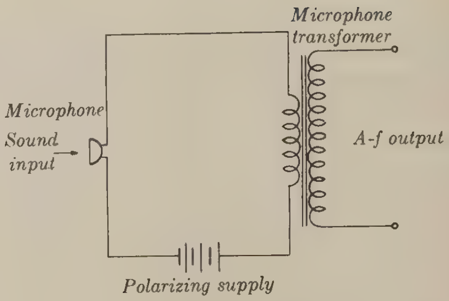


Fig. 8-32 Single-button carbon microphone circuit.

being connected to the metal diaphragm and the other to the metal cup. The carbon button is connected in series with the primary of a step-up transformer and a low-voltage d-c supply of about 2 to 6 volts (Fig. 8-32). When no sound is allowed to enter the microphone, the diaphragm is at rest and the resistance of the carbon button remains constant at about 100 ohms. A steady current therefore flows through the carbon button and primary winding of the microphone transformer. When a compression enters the microphone the diaphragm is pushed in slightly. This presses the carbon granules more closely together and causes their resistance to decrease. The current flowing through the carbon button and primary winding increases and an e.m.f. is induced in the secondary winding of the transformer. When a rarefaction enters the microphone, the diaphragm moves out slightly. This reduces the pressure on the carbon granules, their resistance increases and the primary current decreases. Hence an e.m.f. is induced in the secondary winding in the opposite direction to that caused by a compression. As a result, the sound waves entering the microphone cause the diaphragm to vibrate thus varying the primary current. This in turn produces an alternating e.m.f. in the secondary winding of the transformer. The frequency of the induced e.m.f. is the same as the frequency or pitch of the sound wave. The amplitude of the induced e.m.f. is determined by the intensity of the sound wave.

The carbon-button microphone has a high sensitivity, the output of the microphone transformer having an amplitude up to 10 volts for normal speaking input. However, the fidelity of this type of microphone does not compare favourably with other types so that its use is generally restricted to speech since understandable speech transmission requires only a limited frequency range.

Mechanical resonance of the diaphragm at a frequency in the upper audio range and the sluggishness of the diaphragm contribute to the lack of fidelity. In addition there is a high inherent noise output from carbon-button microphones.

The fidelity of the carbon microphone can be considerably increased by employing two buttons as shown in Fig. 8-33. This type is called the *double-button* microphone. Its construction is similar to that of the single-button type except that a metal cup containing granules is placed on each side of the diaphragm. There are three electrical leads from the microphone, one from each cup and one from the diaphragm. When the diaphragm moves in such a way as to increase the pressure on one button, it reduces the pressure on the other button. Thus, when the current through one button increases, the current through the other button decreases. The micro-

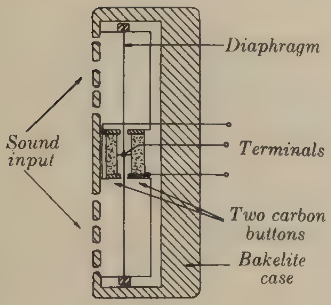


FIG.8-33 Double-button carbon microphone.

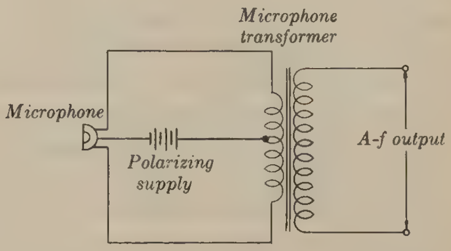


FIG.8-34 Double-button carbon microphone circuit.

phone is connected in a push-pull type of circuit (Fig. 8-34). An increasing current in the clockwise direction in one half of the primary winding produces the same effect in the secondary as a decreasing current in the counterclockwise direction in the other half of the primary. Hence, the e.m.f.'s induced in the secondary circuit by the two primary currents are in phase with each other and add together.

The fidelity of the double-button microphone is higher than that of the single-button type because second harmonic distortion is largely eliminated by the push-pull operation. However the output of this microphone is considerably less than that of the single-button type as less current flows through the double-button microphone. The high noise level is characteristic of all carbon microphones.

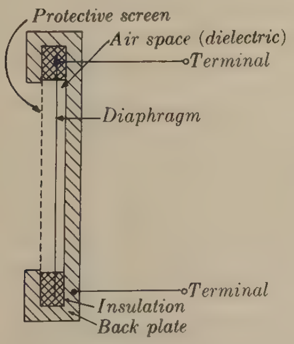


FIG.8-35 Capacitor microphone.

B. Capacitor Microphone

A capacitor microphone is shown in Fig. 8-35. It consists of a very thin, tightly stretched diaphragm of aluminum alloy placed close to a metal back plate. The diaphragm and back plate are insulated from each other and separated by an air space about .001 inch thick, forming a capacitor.

The microphone is connected in series with a d-c supply of about 200 volts and a very high resistance R in the neighbourhood of 20 megohms, as shown in Fig. 8-36. Sound waves falling on the diaphragm cause it to vibrate, varying its capacitance at the a-f rate. When no sound enters the microphone the microphone capacitor is charged to 200 V and there is no voltage across R. When a compression enters the microphone the diaphragm is pushed toward the back plate, the capacitance is increased and the p.d. between the capacitor plates is decreased. The capacitor begins to charge through R producing a voltage across it. When a rarefaction enters the

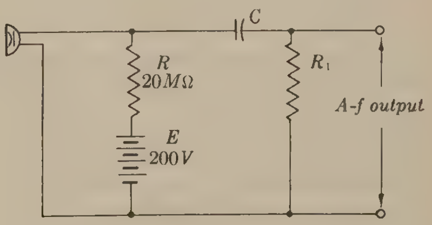


FIG.8-36 Capacitor microphone circuit.

microphone the diaphragm moves out from the back plate and beyond its initial position. The capacitance is decreased below its initial value, the p.d. between the capacitor plates is increased and the capacitor begins to discharge through R producing a voltage in the opposite direction across R. Thus, an audio-frequency voltage is produced across R which is applied via C, across R_1 to the succeeding amplifier stage. C blocks the $200V$ from the next stage. The resistance in the circuit is made very large so that the charge on the capacitor remains substantially constant throughout the a-f cycle. As a result, there is a linear relationship between the amplitude of the a-c voltage across R and the change in capacitance which depends on the sound intensity. The frequency of the a-c voltage across R corresponds to the pitch of the sound.

The sensitivity of the capacitor microphone is very low and it ordinarily requires two additional stages of voltage amplification to raise the level of the output to that of a single-button carbon microphone. No impedance-matching transformer is required between the capacitor microphone and the grid circuit of the first amplifier as the output impedance of the microphone is very high. For this reason the microphone must be well shielded to prevent pickup of a-c hum voltages. The frequency response of the microphone is ordinarily better than that of the carbon microphone and there is no background hiss.

C. *Velocity or Ribbon Microphone*

The velocity microphone shown in Fig. 8-37 contains a thin corrugated metallic ribbon which is suspended between the pole pieces of a permanent horseshoe magnet. The ends of the ribbon are insulated from the magnet and form the terminals of the microphone. Sound waves entering the microphone fall upon the ribbon and set it in vibration. Since the ribbon is a conductor moving across a magnetic field, an e.m.f. is induced in it. The frequency and amplitude of the e.m.f. are determined by the pitch and intensity of the sound as in the case of the carbon and capacitor microphones. The name *velocity microphone* arises from the fact that the movement of the ribbon is proportional to the velocity of the air molecules set in motion by the sound wave.

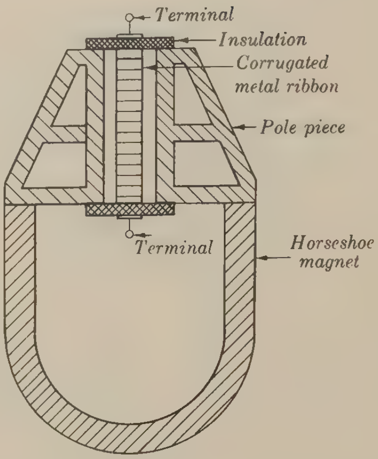


Fig. 8-37 Velocity or ribbon microphone.

The frequency response of the microphone is flat over a wide range of audio frequencies so that its fidelity is comparatively high. The output impedance of the microphone is extremely low, of the order of 1 ohm. It is common practice to use two step-up transformers T_1 and T_2 in cascade (Fig. 8-38) to match the low impedance of the microphone to the input impedance of the amplifier. The first transformer ordinarily matches the ribbon impedance to 200 ohms impedance and the second one matches 200 ohms to the grid circuit impedance. This arrangement overcomes the difficulty of designing a single ribbon-to-grid transformer with satisfactory fidelity characteristics.

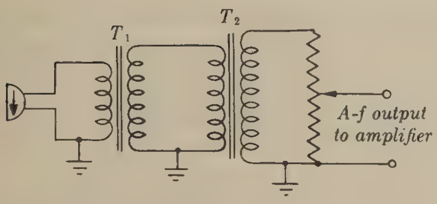


FIG.8-38 Matching impedance of ribbon microphone to impedance of amplifier.

The output voltage from the high-impedance terminals of the second transformer is comparable with that of the capacitor microphone. This is considerably less than the output of the carbon microphone.

D. Dynamic Microphone

The dynamic microphone (Fig. 8-39) is similar in construction to a small permanent-magnet loudspeaker. It consists of a light diaphragm or cone connected to a fine coil of wire which is suspended between the poles of a strong permanent magnet. The sound waves fall on the diaphragm or cone and set it in vibration. This in turn causes the coil to vibrate in the field of the permanent magnet. An alternating e.m.f. is produced across the ends of the coil as in the case of the ribbon microphone. As the output impedance of this microphone is also low, about 30 ohms, a step-up transformer is used to match the impedance to that of the amplifier grid circuit. The output of the dynamic microphone is comparable with that of the capacitor and ribbon microphones.

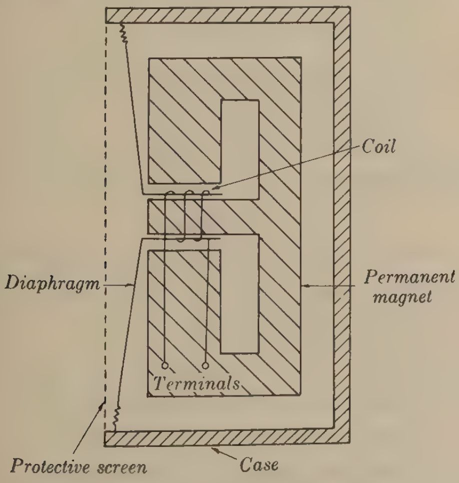


FIG.8-39 Dynamic microphone.

E. Crystal Microphone

The crystal microphone depends for its operation on the piezo-electric effect. That is, certain crystals develop an alternating p.d. between opposite faces when subjected to an alternating pressure. While quartz and Rochelle salts exhibit this effect, the latter crystals are generally employed in microphones because of their greater sensitivity. Two types of crystal microphones are in common use, the *diaphragm* type and the *sound-cell* type.

The diaphragm microphone contains a unit consisting of two crystal slices cemented together in such a way that when one slice tends to expand due to an applied voltage the other tends to contract. A fibre diaphragm is mechanically coupled to the crystal unit. The sound waves entering the microphone cause the diaphragm to vibrate. These vibrations are communicated to the crystal causing it to bend at the same frequency as the sound wave. The bending of the crystals causes an alternating voltage to be developed between its opposite faces. This voltage is picked up by a pair of foil electrodes in contact with the opposite faces of the crystal and provides the a-f output.

The sound-cell type of microphone is made up of a number of small crystal units called *sound cells.* The cells are connected in series or series-parallel to increase the output voltage and to provide a more uniform frequency response. Each sound cell consists of two pairs of crystal plates, the pairs being separated from each other by a small air space (Fig. 8-40). The individual

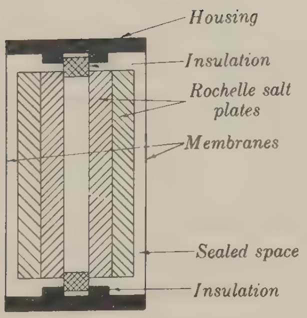


FIG.8-40 Sound cell of crystal microphone.

plates are sections of Rochelle salt crystals about $\frac{1}{2}$ inch square and $\frac{1}{5}$ inch thick. The plates of each pair are cemented together back to back as in the case of the unit in the diaphragm microphone. The crystal plates are enclosed in an airtight compartment. The alternating pressure in the sound wave is applied directly to the individual sound cells. A protective membrane is used in some types of crystal microphones.

The output impedance of the crystal microphone is extremely high and is capacitive. It is therefore connected directly to the input circuit of the amplifier using a grid-leak resistor of about 5 megohms. The voltage output of the crystal microphone is slightly greater than that of the capacitor and velocity types. The fidelity of the microphone is quite high, especially with the sound-cell type. A uniform frequency response from 100 to 5000 cycles per second is usual with a well designed crystal microphone.

8.17 Questions and problems.

1. What is the purpose of a radio transmitter?
2. State the essential parts of a transmitter.
3. Explain the disadvantages of a transmitter consisting of an oscillator, power supply and aerial only.
4. Give reasons why high frequency stability is required in a transmitter.
5. Explain the term frequency tolerance.
6. How is the frequency tolerance of ship transmitters checked?
7. Define the term frequency drift.
8. How is frequency drift avoided?
9. Describe the construction and action of a negative-temperature-coefficient capacitor and an anti-drift coil.
10. Why is loose coupling to the oscillator desirable?
11. Explain the purpose of a buffer amplifier.
12. State any other causes of frequency instability in a transmitter.
13. Explain how the effect of each of the above is minimized.
14. Why is it not desirable to key the oscillator?
15. Explain the construction of a master oscillator.
16. Compare m.o. control and crystal control as regards frequency, power, frequency stability and tuning flexibility.
17. Why must frequency multipliers be used at high frequencies with crystal control?
18. Draw a block diagram of a c-w transmitter with m.o. control.
19. Draw a block diagram of an m.c.w. transmitter with crystal control and grid modulation.
20. Draw a block diagram of an r-t transmitter with m.o. and crystal control and anode modulation.
21. What advantages result when both m.o. and crystal control are employed?

22. Explain the term paralyzing bias.
23. Draw a block diagram of a transmitter for c-w, m-c-w or r-t transmission with m.o. and crystal control and grid modulation.
24. Explain the action of each stage of the above transmitter.
25. Explain the purpose and principle of operation of a frequency multiplier.
26. Draw a circuit diagram of a frequency multiplier stage.
27. Explain the action of the above circuit.
28. Why is the efficiency of a frequency multiplier less when it operates as a tripler than as a doubler?
29. What efficiency is expected from a doubler if its efficiency as an amplifier at the fundamental frequency is 80%?
30. What is meant by optimum angle of flow in a multiplier? What should its value be for a doubler?
31. Draw a diagram to illustrate the input voltage, anode current and output voltage waveforms of a frequency doubler with an operating angle of 90°.
32. Why is a high-Q tuned circuit desirable in a multiplier?
33. Why can a push-pull stage not be operated as a doubler?
34. Draw and explain the action of a modified push-pull circuit which may be used as a doubler.
35. Why is neutralization necessary in a triode amplifier?
36. Why are a-f amplifiers not usually neutralized?
37. Why are pentodes not used in place of triodes as r-f power amplifiers to avoid neutralizing arrangements?
38. Draw the circuit of a grid-neutralized amplifier.
39. Explain the action of a grid-neutralized amplifier.
40. What is the condition for complete neutralization?
41. Draw the circuit of an anode-neutralized amplifier.
42. Explain its action using the equivalent bridge circuit.
43. Draw a circuit for neutralization at one frequency only.
44. Draw the circuit of a neutralized push-pull amplifier.
45. Why must the neutralizing capacitor have a high-voltage rating?
46. Upon what principle does the neutralizing procedure depend?
47. Describe the neutralizing procedure for the circuit of question 41.
48. Why is a c-w signal employed during the neutralization procedure?
49. Why is an amplifier usually neutralized on its highest range?
50. Explain the neutralization procedure for a push-pull amplifier.
51. What are parasitic oscillations?
52. Why are parasitic oscillations not desirable?
53. Explain how h-f parasitics occur.
54. Explain how h-f parasitics are eliminated.
55. How are l-f parasitics set up?
56. How are l-f parasitics eliminated?
57. What power sources are likely to be available in a ship for the transmitter equipments?
58. What kind of power supply is required as the actual input to a transmitter?
59. What type of power supply is required in a high-power transmitter?
60. What are the control circuits?

61. Explain the positions and purpose of the safety door and panel switches.

62. Explain the difference between stand-by and normal operation.

63. Explain the purpose and action of
    (a) break-in keying relay
    (b) overload relay
    (c) filament time-delay relay.

64. Explain the purpose of a wavemeter or frequency meter.

65. Give the formula which relates frequency and wavelength.

66. A wavemeter gives the following frequency values for three transmitters. Calculate the wavelength in metres in each case.
    (a) 500 Kc/s
    (b) 12.5 Mc/s
    (c) 90 Mc/s.

67. The frequency of a transmitter is measured as 1.85 Mc/s. Calculate the corresponding wavelength in metres.

68. Calculate the frequencies of transmission for the following wavelengths:
    (a) 5 metres
    (b) 1.5 metres
    (c) 10 centimetres.

69. What are the essential parts of an absorption wavemeter?

70. Explain the action of a simple absorption wavemeter.

71. Describe three methods of resonance indication which are employed with absorption wavemeters.

72. Draw the circuit of an absorption meter with a vaccum-tube voltmeter suitable for measuring frequency over several ranges.

73. Explain the action of the above circuit.

74. Why must the position of the wavemeter be fixed when a measurement is made?

75. When and why are frequency calibration charts used?

76. What are the essential parts of a heterodyne wavemeter?

77. Explain the principle of a heterodyne wavemeter.

78. Draw and explain the action of a simple heterodyne wavemeter circuit.

79. What is the method of resonance indication in a heterodyne wavemeter?

80. List the preliminary adjustments necessary in tuning up a transmitter.

81. Explain the purpose of each adjustment.

82. Explain the tuning procedure for a c-w transmitter having the following stages
    (a) crystal oscillator-master oscillator
    (b) frequency doubler
    (c) driver
    (d) power amplifier
    (e) transformer-coupled aerial.

83. What resonance indications are used in tuning an amplifier to a given r-f signal?

84. Why is a crystal oscillator detuned slightly?

85. On which side of resonance is a Pierce crystal oscillator detuned? A Miller crystal oscillator?

86. What type of aerial is generally used with r-f transmitters?

87. Why must this aerial be inductively loaded?

88. What is the object in tuning the aerial?

89. By means of sketches illustrate several methods of coupling a power amplifier to an aerial.

90. Explain how the aerial is tuned in each method.

91. A power amplifier is transformer coupled to a quarter-wave Marconi aerial having rough and fine inductance tuning. Make a circuit diagram and explain the aerial tuning procedure.

92. What condition indicates that the aerial is tuned?

93. What condition indicates that the aerial is matched to the p.a.?

94. Why is matching necessary between a power amplifier and an aerial?

95. After an aerial tuning procedure has been carried out it is found that a slight change in the capacitance of the power amplifier tank causes an increase in aerial current. What condition is suggested by this observation? What corrective steps should be taken?

96. Draw a block diagram of the transmitter shown in Fig. 8.19.

97. Draw the m.o. circuit separately, name the type of oscillator and explain the purpose of L_1, C_{17} and C_{18}, C_2, C_{37} and R_{13}, and C_3.

98. Draw the crystal oscillator circuit, name the type of oscillator and explain the purpose of each component.

99. Draw the V_2 circuit as it appears when operating as an untuned amplifier. Name the type of circuit and explain the purpose of C_{26} and C_{27}, C_{22} and R_5.

100. Draw the V_2 circuit as it appears when operating as a tripler. Name the type of circuit and explain its action. Explain the purpose of each component.

101. Draw the driver circuit, name the type of circuit and explain the purpose of C_{15}, C_{16}, L_{10}, R_9, R_{10} and C_{47}, C_{30}.

102. Draw the power amplifier and aerial circuit for c-w operation. Name the type of circuit and explain the purpose of L_{15}, C_{51}, R_{18}, R_{19} and L_{20}.

103. Explain how the paralyzing bias is obtained when the key is open.

104. (a) What are the objections to keying an oscillator stage?
    (b) How are these objections overcome in this transmitter?

105. Explain the sequence of events when the press-to-talk switch or key is closed assuming the transmitter to be ready to operate.

106. (a) Explain the time-delay action.
    (b) How may this delay be reduced?

107. (a) Draw the power-amplifier grid circuit and the modulator for (i) m-c-w (ii) r-t.
    (b) Name the type of circuit in each case.
    (c) Explain the purpose of T_1, R_{31}, T_2, C_{45} and R_{21}, J_1, R_{30} and R_{29}, S_2 and R_{26}.

108. Explain the action of the break-in keying relay E_4.

109. What preliminary adjustments are made before tuning up this transmitter?

110. Explain the tuning procedure for the transmitter stage by stage.

111. Describe the construction of a single-button carbon microphone.

112. What is the purpose of the d-c supply connected in the carbon microphone circuit?

113. Explain how sound waves entering the microphone are made to control the flow of current from the d-c supply.

114. What determines the frequency of the a-c output of the microphone? What determines the amplitude of the output?

115. What is the purpose of the transformer connected between the microphone and the amplifier?

116. What are the chief advantages of a carbon microphone as compared with other types?

117. What is the chief disadvantage of a single-button carbon microphone?

118. Make a diagram of a double-button carbon microphone and its associated circuit including the transformer.

119. Describe the operation of the circuit above.

120. In what respect is the double-button carbon microphone inferior to the single-button type?

121. How does the fidelity of the double-button microphone compare with that of the single-button microphone?

122. Account for this difference in fidelity.

123. Make a cross-section diagram of a capacitor microphone.

124. Explain the operation of the capacitor microphone.

125. Account for the very high value of resistance connected in the output circuit of this type of microphone.

126. How does the impedance of the capacitor microphone compare with that of the carbon microphone?

127. What is an alternative name for the ribbon microphone? Explain the significance of each name.

128. Explain the operation of the ribbon microphone when sound waves are made to actuate the ribbon.

129. Explain the purpose of T_1 and T_2 in Fig. 8-38.

130. How does the sensitivity of the ribbon microphone compare with the sensitivity of the carbon and capacitor microphones?

131. A permanent-magnet speaker is used as a microphone. What name is given to this type of microphone? In what main respect does it differ from the ribbon microphone?

132. Name two types of crystal microphones.

133. What type of crystal is generally used in crystal microphones?

134. Explain how the diaphragm type of crystal microphone generates an a-c voltage.

135. Describe the construction of a sound cell.

136. State two advantages of the sound-cell microphone over the diaphragm type.

CHAPTER IX

RADIO COMMUNICATION RECEPTION

9.1 Principles of reception. The essential components of a receiving station in a radio communication system are shown in Fig. 9-1. The purpose of the receiver is to receive the signal sent out by the transmitting station and to reproduce the intelligence which the sender desires to communicate.

A. Aerial Circuit

The receiving aerial absorbs energy from passing radio waves and feeds it to the receiver. This energy transfer occurs when the electromagnetic radio waves from the transmitter cut through the receiving aerial and induce in it an alternating current. This alternating current is similar to that in the transmitting aerial but greatly reduced in amplitude. Each of the transmitting stations, within range of the receiver, induces a current in the aerial at the station's own frequency. These currents flow simultaneously in the aerial circuit and are selected one at a time by the tuner circuit.

B. Tuner

The tuning circuit of the receiver is a resonant circuit consisting of an inductor and capacitor connected in series. It is an acceptor circuit which receives energy from the currents induced in the receiving aerial. As in the case of the aerial circuit, many circulating currents normally flow in the tuner circuit simultaneously, each current flowing at the frequency of the transmitter signal producing it. The capacitance of the resonant circuit is then altered to tune the circuit to resonance at the frequency which it is desired to select. A comparatively large current circulates in the tuner at this resonant frequency while the currents at frequencies off resonance are very small. This selected current produces a correspondingly large alternating r-f voltage across the tuning capacitor.

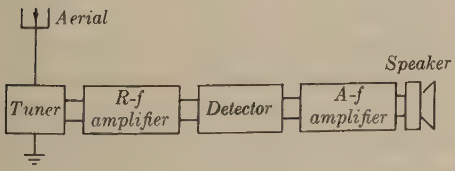


Fig.9-1 Block diagram of a simple receiver.

C. Radio-Frequency Amplifier

The r-f voltage produced in the tuner circuit, even at the resonant frequency, is so small that it usually requires amplification prior to the detector stage. The radio-frequency amplifier consists of one or more stages of voltage amplifiers which increase the amplitude of the signal voltage. This amplifier may also contain resonant circuits which are tuned to the frequency of the signal by the same control which operates the aerial tuning circuit. This provides a greater degree of selectivity than that given by a single tuning stage.

D. Detector

The purpose of the detector circuit is to recover from the radio-frequency signal the audio intelligence supplied at the transmitter. The type of detector may vary considerably according to the power level at which the signal is detected. Diode detection is the most common type; crystal detection, grid

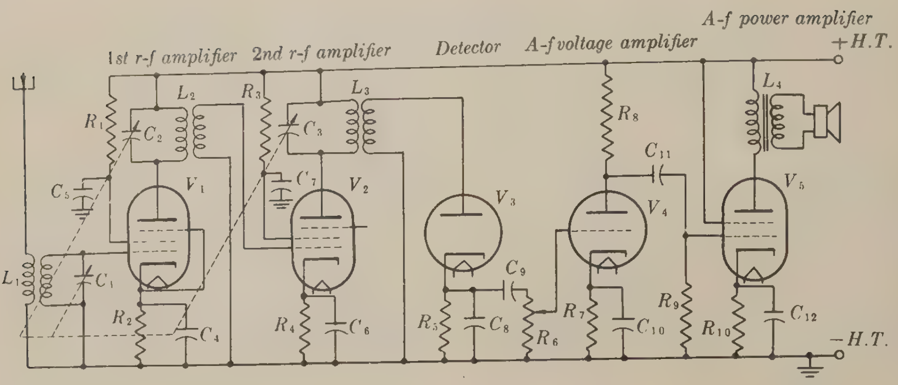


FIG. 9-2 A simple t-r-f receiver.

detection and anode detection are rarely used in Service equipment. Various types of detectors are described in Chapter VII. The output of the detector is the audio signal voltage which forms the input to the audio-frequency amplifier.

E. Audio-Frequency Amplifier

The audio-frequency section consists of two types of amplifiers, a voltage amplifier and a power amplifier. The a-f voltage amplifier usually employs a high-gain triode or pentode valve in a resistance-capacitance coupled circuit. It increases the amplitude of the a-f signal to the level required by the power amplifier. The power amplifier circuit contains a triode or beam power valve which provides sufficient a-f current to operate headphones or a loudspeaker.

9.2 A simple t-r-f receiver. The circuit diagram of a simple tuned-radio-frequency (t-r-f) receiver is shown in Fig. 9-2. It consists of two r-f amplifiers, a diode detector, an a-f voltage amplifier and an a-f power amplifier. The aerial circuit is coupled to the tuner by the aerial transformer L_1. The tuning capacitor C_1 is one section of a three-gang capacitor, of which the other two sections are C_2 and C_3. Pentode valves are used in the r-f stages because they are more stable than triodes in high-frequency circuits. The screen grid construction reduces the grid-anode capacitance and thereby reduces the tendency to produce oscillations. R_1 is the screen voltage dropping resistor and C_5 the screen by-pass capacitor. C_4 and R_2 provide cathode bias for Class A operation of the amplifier.

Radio waves from the transmitter passing the aerial wire causes an alternating current to flow in the aerial-ground circuit at the transmitter frequency. This current flowing in the primary winding of L_1 induces a similar current in the secondary circuit of L_1 which sets up a circulating current in the tuner. C_1 is then adjusted so that the tuner is in resonance with the transmitter frequency and the circulating current in L_1C_1 becomes a maximum at this frequency. The alternating voltage developed across C_1 by this circulating current is applied between grid and cathode of V_1 to control the flow of current through the valve. The alternating component of the current through the valve causes a circulating current to build up in the tuned anode circuit L_2C_2. The tuning capacitor C_2 is operated by the same control as that on C_1 so that the two circuits are always in resonance at the same frequency. The r-f current which circulates in L_2C_2 is greater in amplitude than that in L_1C_1 because of the amplification of the circuit. The output of V_1 is coupled through the secondary winding of the transformer L_2 to the grid circuit of V_2. The operation of V_2 is similar to that of V_1 but with an amplified signal. Waveforms of input and output voltages of V_1 and V_2 are shown in Fig. 9-3 (a), (b), (c) for a modulated continuous wave.

The output of V_2 is transformer coupled to the detector stage. This circuit consists of a diode valve which rectifies the r-f signal and a filter circuit R_5C_8. The r-f component of rectified current which flows through V_3 is by-passed by C_8. The audio-frequency component of the rectified current flows through R_5 and develops the a-f signal voltage across this resistor. Waveforms of input voltage, rectified current and output voltage of V_3 are shown in Fig. 9-3 (c), (d) and (e).

The a-f signal voltage is capacitively coupled through C_9 to the potentiometer R_6. The potentiometer is the volume control for the receiver. The entire a-f signal voltage from the detector is applied across R_6 and that part of the voltage between the sliding contact and ground is used as the input to V_4. By moving the sliding contact up on R_6, the amplitude of the voltage to V_4 is increased and hence the volume of the final output is increased.

The a-f voltage amplifier is a Class A resistance-capacitance coupled circuit. The valve V_4 is a high-mu triode. Bias is provided by R_7C_{10} in the cathode circuit and the output voltage is developed across the load resistor R_8. Waveforms of the input and output voltages of this stage are shown in Fig. 9-3 (e) and (f).

The power amplifier stage V_5 is a beam-power amplifier which produces an audio-frequency current of large amplitude for operating the loudspeaker. The fluctuations in current through this valve may be as much as 1000 times the corresponding variations of current through the previous valve. Bias is provided by $R_{10}C_{12}$ in the cathode circuit for Class A operation. The output transformer L_4 is used to match the high impedance of the valve circuit to the low impedance of the speaker. The relative impedances of these two circuits is frequently of the order of 900:1 which requires a step-down transformer of 30:1 to provide maximum power to the speaker. Such a voltage reduction provides a corresponding step-up in current to the speaker, which is a current-operated device. Waveforms of the input and output voltages of the power stage are shown in Fig. 9-3 (f) and (g).

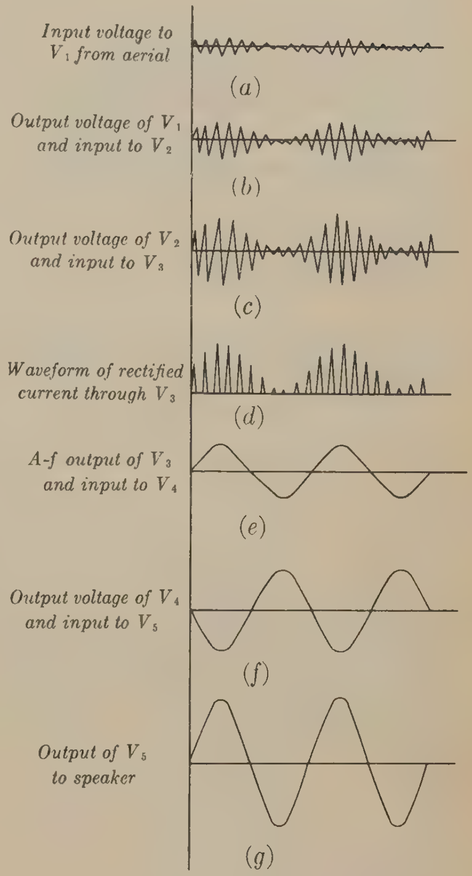

Input voltage to V_1 from aerial
(a)
Output voltage of V_1 and input to V_2
(b)
Output voltage of V_2 and input to V_3
(c)
Waveform of rectified current through V_3
(d)
A-f output of V_3 and input to V_4
(e)
Output voltage of V_4 and input to V_5
(f)
Output of V_5 to speaker
(g)

FIG. 9-3 Input and output voltages for the t-r-f receiver stages.

The receiver just described is simple in construction and in theory of operation. While it may prove satisfactory for reception from a nearby transmitter of considerable power, it is generally inadequate for long-distance communication for a number of reasons. The requirements of a good communications receiver are discussed in the following section.

9.3 Requirements of a communications receiver.

A. Selectivity

The receiving aerial comes under the influence of a large number of transmissions, each of which induces in the aerial an e.m.f. at its own carrier frequency. To avoid interference from adjacent channels, circuits selective to only a small band of frequencies are employed. The degree to which the receiver is capable of differentiating between the desired signal and signals of other carrier frequencies is called the *selectivity* of the receiver. This selection of frequency occurs mainly in the tuner circuit of the first stage. However, each resonant or tuned circuit in the receiver contributes to this effect and the overall selectivity depends on the selectivity of every circuit. A high degree of selectivity requires that the tuner have a low resistance, high L/C ratio and that it be loosely coupled to the aerial.

B. Sensitivity

The aerial voltages induced by the electromagnetic waves are so small that a large amount of amplification is necessary for an audible output. This amplification takes place in most of the stages of the receiver. The *sensitivity* of a receiver is a measure of the overall amplification. It is defined as the number of microvolts input r-f, modulated 30% at 400 c/s, required to produce an audio output of 500 milliwatts. A good receiver will have a sensitivity of 1 to 10 microvolts. Increasing the sensitivity beyond this point increases the volume of incidental aerial and valve noise in most receivers to the point where it obscures the signal.

C. Signal-To-Noise Ratio

The *signal-to-noise* ratio is the ratio of audio-signal power output to noise power output at any particular frequency. Motors, generators and other electrical equipment cause noise voltages to be present in the aerial at all frequencies. The electrical circuits and valves of any receiver also have a certain amount of inherent noise caused by the random movement of electrons. The amount of noise is directly proportional to the bandpass of the receiver. Thus, a receiver with good selectivity will have a better signal-to-noise ratio and better sensitivity than a receiver whose selectivity is poor.

Some receivers have a variable bandpass which allows the operator to adjust the sensitivity of the set to obtain a good signal-to-noise ratio. A better signal-to-noise ratio is also obtained by making the aerial directive and pointing it at the transmitter, by making the transmitter aerial directive and pointing it at the receiver, or both. In each case the signal strength at the receiver is increased, making the signal-to-noise ratio greater.

D. Fidelity

Fidelity is the degree to which the receiver faithfully reproduces the modulation of the incoming r-f carrier. For high fidelity the bandwidth of the receiver must be wide enough to include the extreme sideband frequencies. In the broadcast band, the bandwidth is usually 5 Kc/s on either side of the carrier, making 10 Kc/s altogether. However since increasing the bandwidth to obtain good fidelity decreases the sensitivity of a receiver, a compromise is made between good sensitivity and good fidelity.

E. Stability

The ability of a receiver to give a constant output from a signal of constant strength and frequency is known as its *stability*. The frequency to which the receiver is tuned may be altered by changes in temperature or in power supply voltage. Negative-temperature-coefficient capacitors are used to eliminate the effect of temperature changes and well regulated power supplies eliminate possible changes in H.T. voltage.

9.4 Types of r-f receiver signals. The simple t-r-f receiver shown in Fig. 9-2 is suitable for the reception of radio waves which are modulated, in amplitude at the transmitter. Two types of amplitude-modulated waves, together with a wave of constant amplitude are shown in Fig. 9-4. A brief comparison of the advantages and disadvantages of these three types of radio communication is given in the following section. So far as transmission is concerned, they are described in the order of increasing complexity, which is the order of simplicity so far as reception and de-coding are concerned.

The *c-w signal*, Fig. 9-4 (a), consists of a radio-frequency carrier wave of constant amplitude which is keyed by the operator at the transmitter. Wave trains of short and long durations separated by spaces of no transmission represent the dots, dashes and spaces which constitute the Morse code. Transmission by this method requires only a very narrow band of frequencies so that a great many channels may be accommodated in a relatively narrow band of allocated radio frequencies. It has the disadvantage however that the signal must be modulated at the receiver before an audible output is produced.

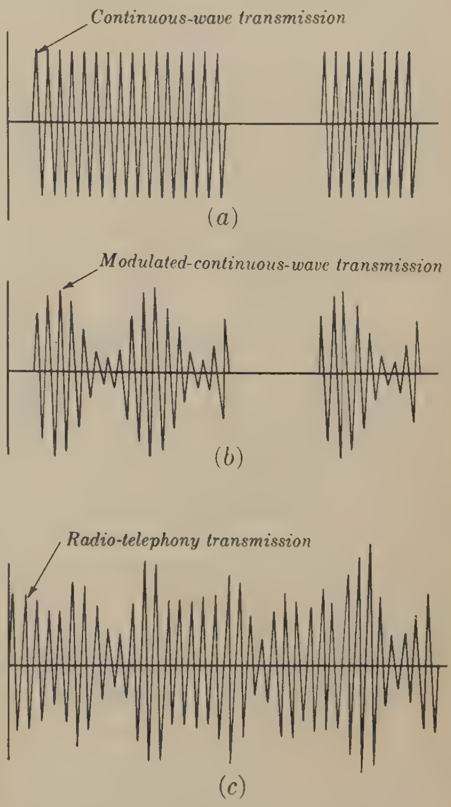

Continuous-wave transmission
(a)
Modulated-continuous-wave transmission
(b)
Radio-telephony transmission
(c)

FIG.9-4 Types of r-f receiver signals.

The *m-c-w signal*, Fig. 9-4 (b), consists of an r-f carrier modulated by a single-frequency audio signal produced at the transmitter. The transmitter is keyed for Morse code transmission. The output of the detector stage in the receiver is then an audible signal. Such a system requires a wider transmission bandwidth than the c-w system. The bandwidth extends above and below the carrier frequency by an amount equal to the modulating frequency. Thus with *m-c-w* transmission fewer channels may be accommodated in a given range of frequencies than with c-w transmission.

The waveform of an *r-t signal* is shown in Fig. 9-4 (c). The modulation envelope is a complex waveform resulting from many audio-frequency currents of different amplitudes produced by the microphone at the transmitter. Consequently the bandwidth required to transmit such an r-t signal with its many r-f frequencies is much wider than for either c-w or m-c-w transmission. In the broadcast band, a range of 10 Kc/s is allocated to each broadcasting station. This permits the transmission of sideband frequencies up to 5 Kc/s above and 5 Kc/s below the carrier frequency. The use of r-t must therefore be restricted to bands where a great deal of traffic must be passed and the number of channels in use must be kept to a maximum.

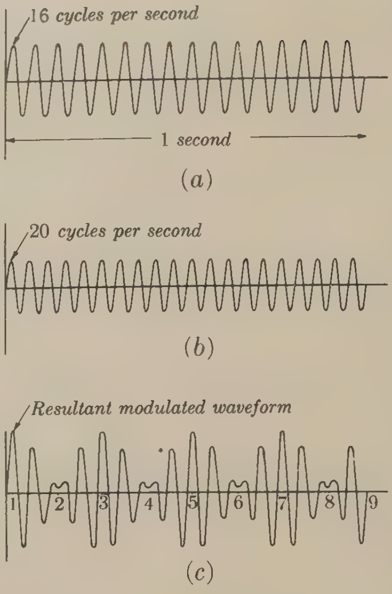


Fig.9-5 Waveforms illustrating the heterodyne principle.

9.5 The heterodyne principle. It was pointed out in the previous section that in order to produce an audible output from a c-w signal the receiver must modulate the signal. This is accomplished by employing the *heterodyne principle.* If two continuous waves of different frequencies are applied to the same circuit, the result is a new wave whose amplitude is modulated. The frequency of the modulation envelope is equal to the difference of the two frequencies which were mixed. This is called the heterodyne principle and is applicable at audio frequencies as well as radio frequencies. It may be demonstrated with sound waves as well as radio waves. For simplicity in explanation, consider a wave of 16 cycles per second to be mixed with a wave of 20 cycles per second as shown in Fig. 9-5 (a) and (b). The resultant wave is the modulated one shown in Fig. 9-5 (c). At the instants indicated by 1, 3, 5, etc., the 16 and 20 c/s waves are in the same phase. The resultant wave, therefore, reaches its maximum amplitude which is equal to the sum of the amplitudes of the two original waves. At instants 2, 4, 6, etc., the waves are in anti-phase and tend to cancel each other. The resultant wave then falls to a minimum amplitude which equals the difference of the amplitudes of the original waves. It may be seen from the diagram that the number of times the waves are in phase in one second is equal to the difference of their frequencies. Each complete cycle of amplitude variations is called a *beat.* The number of beats produced each second is called the *beat frequency.* The beat frequency therefore equals the difference between the frequencies of the two mixed signals. If a continuous wave is to be modulated at an audio-frequency rate, a second c-w, produced at the receiver and differing from the received signal by a few hundred or a few thousand cycles per second, is injected into the circuit prior to the detector. The modulation frequency and hence the tone of the audio output may be varied by adjusting the frequency of the injected signal. The circuit which produces this r-f signal is called the *beat-frequency oscillator.* A typical beat-frequency oscillator circuit that may be used with the simple t-r-f receiver of Fig. 9-2 is shown in Fig. 9-6. It consists of a triode valve in a modified Hartley oscillator circuit. The tuned circuit L_1C_2 is designed to produce oscillations

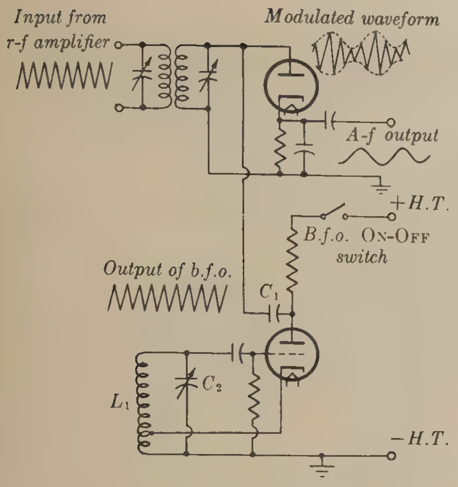


Fig.9-6 A b.f.o. and diode detector circuit.

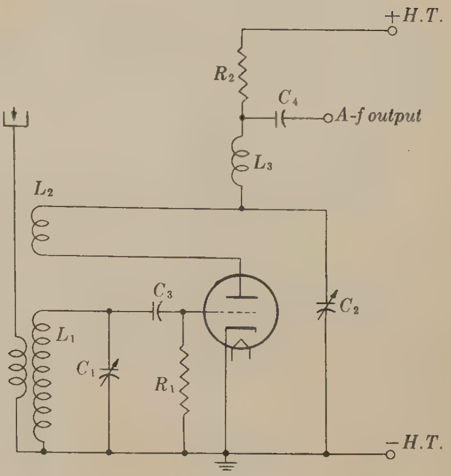


FIG.9-7 Regenerative detector.

over the range of frequencies covered by the main tuning capacitor. The output of the b.f.o. is a continuous wave which is capacitively coupled by C_1 to the anode of the diode detector. Thus the diode is driven at two frequencies, that of the received signal and that of the b.f.o. The result is a modulated wave which is rectified and filtered by the detector circuit. The capacitor C_2 is adjusted so that the output of the b.f.o. differs in frequency from the received signal by an amount equal to the audio frequency desired in the output. The b.f.o. may be tuned slightly above or slightly below the signal frequency. For example, if the signal frequency is 400 Kc/s and the desired audio-frequency note is 1000 c/s, the b.f.o. may be tuned to produce an output of 401 Kc/s or 399 Kc/s. If the b.f.o. is tuned to produce the same frequency as the incoming signal, the received signal remains unmodulated and the result is referred to as *zero beat* or *dead beat*. This may be of use in eliminating an interfering signal. For example, suppose that the desired signal is at 270 Kc/s and an interfering signal is at 269.2 Kc/s. The latter is too close to the desired signal to be tuned out by the tuner circuit. The b.f.o. may be adjusted to produce 269.2 Kc/s. The result is to produce zero beat with the unwanted signal and 800 c/s beat with the desired signal.

The b.f.o. may also be used to eliminate an interfering signal by over-reading. In the example given above, the b.f.o. may be tuned to 270.8 Kc/s. The desired signal then produces an 800 c/s note while the interfering signal produces a 1,600 c/s note. Although both signals are heard, the one note is sufficiently different in pitch from the other that the undesired one can be over-read.

9.6 Regenerative detection. Both the sensitivity and selectivity of a receiver may be increased by the use of a regenerative detector. A detector of this type is shown in Fig. 9-7. The circuit is essentially a grid-leak detector. It also resembles a tuned-grid oscillator in construction and operation. A tuned *LC* circuit is connected between grid and cathode of a triode valve. This circuit is coupled to the aerial and is tuned to the frequency of the signal to be received. Radio waves passing the aerial set up oscillations in the tuned grid circuit as in an r-f amplifier. The r-f voltage de-

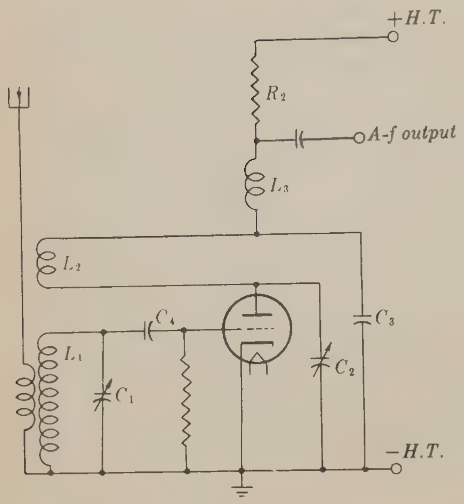


FIG.9-8 Regenerative detector with feedback controlled by anode-to-ground capacitor.

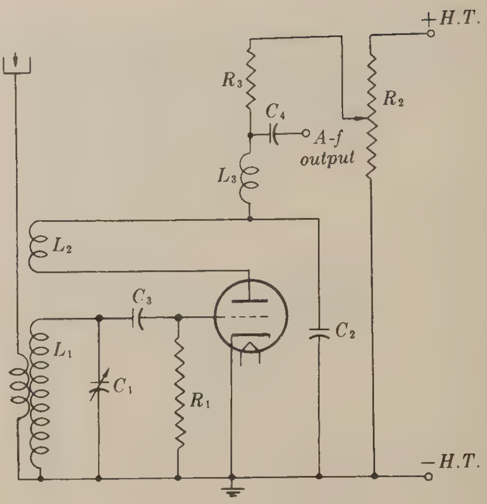


FIG.9-9 Regenerative detector with feedback controlled by anode voltage.

veloped across the tuning capacitor is applied between grid and cathode of the valve. The alternating grid voltage causes an r-f component of current to flow from cathode to anode of the valve. This current flows through a tickler or feedback coil L_2 which is inductively coupled to the grid circuit. The tickler coil is wound in such a way as to increase the amplitude of oscillations in the grid circuit. In this respect the circuit is similar to an oscillator. It differs from an oscillator, however, in that the amount of feedback is not sufficient to maintain oscillation if the signal from the aerial is cut off. Since a regenerative circuit produces a relatively large oscillating current from a small aerial voltage, it has the effect of reducing the resistance of the input circuit. The Q of the circuit is therefore increased and so the selectivity is also increased. It is to be emphasized that the frequency of the oscillations is determined by the aerial signal and that the amplitude rises and falls to follow the modulation envelope of the signal. The regeneration merely increases the degree to which the amplitude of the oscillations rises and falls.

Detection takes place in this circuit as in the simpler grid-leak detector. The r-f component of current flowing through the valve is filtered from the output by the r-f choke L_3 and by-pass capacitor C_2. The a-f output voltage is developed across the anode load resistor R_2.

In order to take advantage of the maximum selectivity and sensitivity available with the regenerative circuit in the reception of signals of various strengths, provision is usually made to vary the amount of regeneration. This is accomplished in the circuit of Fig. 9-7 by the variable capacitor C_2 which completes the path for r-f current through the valve and L_2. If the capacitance of C_2 is small, its reactance is large and there is a high impedance to the flow of r-f current through L_2. Consequently the feedback is reduced and the tendency to produce self-oscillation is lessened. If C_2 is made large, the r-f energy fed back from the anode to the grid circuit is increased and so the regeneration is increased.

In using the regenerative detector for the reception of amplitude-modulated signals, the circuit must be prevented

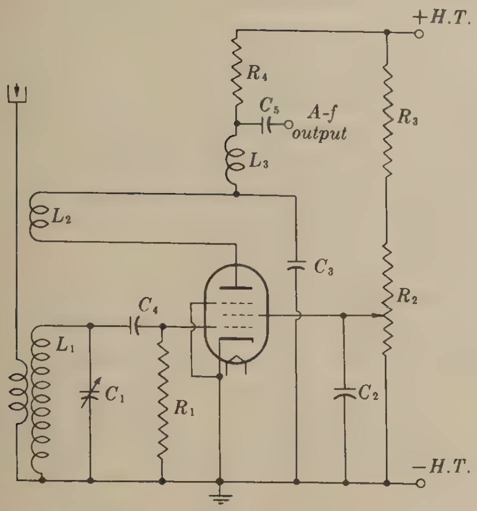


FIG. 9-10 Regenerative detector with feedback from anode to grid circuit controlled by screen voltage.

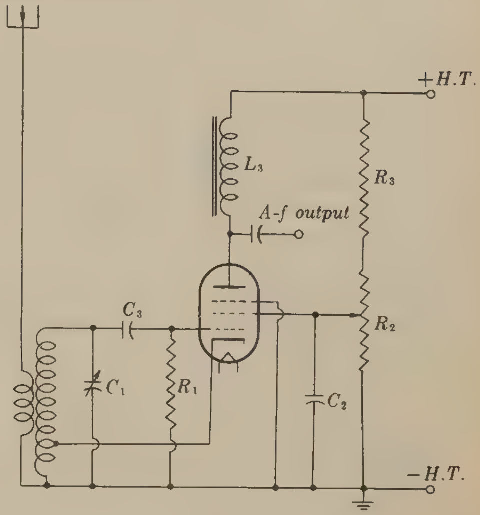


FIG. 9-11 Regenerative detector with feedback in grid-cathode circuit controlled by screen voltage.

from oscillating when no signal is received in the aerial. The regeneration is advanced to the point where self-oscillation begins and is then reduced slightly. Self-oscillation causes howling in the speaker or headphones as the frequency of oscillation is brought close to the frequency of the received signal. The heterodyning effect causes modulation of the signal at a frequency equal to the difference of the heterodyning frequencies. The modulation frequency becomes the output signal of the detector. If this signal is in the a-f range it produces the howling signal in the speaker characteristic of self-oscillation.

Several methods of controlling the amount of regeneration are shown in Figs. 9-8, 9-9, 9-10 and 9-11. In the first of these, feedback is controlled by a variable capacitor C_2 connected between anode and ground. If the capacitance is increased, the r-f component of current through the valve is short-circuited to ground without passing through the feedback coil. The regeneration is thereby reduced.

In the circuit of Fig. 9-9 the amount of regeneration is varied by changing the anode voltage on the valve. If the anode voltage is increased by moving the variable tap of R_2 upward, the current through the valve increases and so the regeneration increases.

A convenient method of controlling feedback with a pentode valve is provided by varying the d-c potential on the screen grid. Two variations of this method are shown in Fig. 9-10 and Fig. 9-11. In either case, an increase in screen potential causes an increase in the gain of the valve and produces more regeneration. Feedback in the circuit of Fig. 9-11 is provided by the cathode circuit as in the case of a series-feed Hartley oscillator. A slight variation of these circuits is shown in Fig. 9-12 in which a tickler coil is placed in the screen circuit instead of the anode circuit. The functions of the various circuit components are the same as those described in the circuits employing triode valves.

9.7 The autodyne. It is pointed out in the previous section that self-oscillation must be avoided in the regenerative detector when receiving modulated sig-

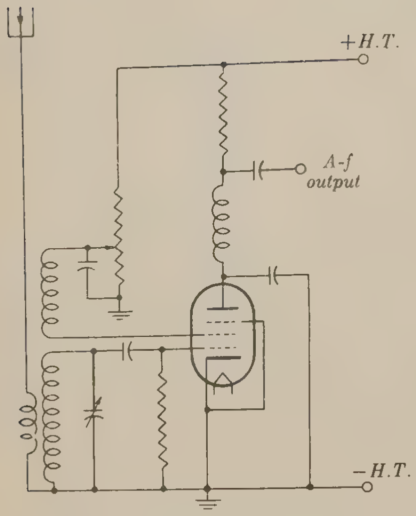


FIG.9-12 Regenerative detector with feedback via the screen circuit.

nals. This type of detector is, however, well suited to the reception of c-w signals but its mode of operation is somewhat different from that just described. In this case the grid circuit is detuned from the aerial signal frequency by a few hundred cycles per second and the regeneration control is advanced sufficiently to produce self-oscillation. The r-f signal from the aerial then mixes with the r-f oscillation produced by the oscillator to produce a modulated signal in the tuned grid circuit. This new signal is detected by the grid-leak detector. The output is an a-f signal whose frequency is equal to the difference between the aerial signal and the self-oscillation frequencies. The tone of the output may be altered by changing the oscillation frequency slightly. For example, if coded c-w signals are received at a frequency of 450 Kc/s, the circuit may be made to oscillate at 450.5 Kc/s in order to produce an a-f signal of .5 Kc/s or 500 c/s. The slight mis-tuning of the grid circuit does not reduce the amplitude of the received signal substantially. The circuit when used in this manner for the reception of c-w signals is called an *autodyne.* The one-valve circuit performs the functions of an r-f amplifier, a beat-frequency oscillator and a detector. Any of the regenerative detectors described in the previous section may be used as an autodyne provided that the regeneration may be advanced sufficiently to produce self-oscillation.

9.8 Super-regenerative detectors. The use of regeneration to increase the a-f output of a detector receiving a small r-f signal is described in the previous sections. Feedback in the circuit makes it possible for a small signal voltage to produce a larger r-f current than a similar non-regenerative circuit would produce. This is equivalent to reducing the effective resistance of the input circuit. The greater the regeneration the less the resistance of this circuit becomes and the more sensitive the receiver is to aerial signals. If the regeneration is advanced sufficiently the circuit breaks into oscillation. It then produces its own r-f current with no applied signal voltage and so the effective resistance of the input circuit is zero or even negative. The r-f oscillation generated by the circuit obscures the aerial signal except in special cases such as the autodyne circuit used for receiving c-w signals.

The *super-regenerative detector* makes use of the fact that maximum sensitivity is obtained when the circuit is on the point of oscillation. Oscillation is permitted to take place but the degree to which it may proceed is controlled so that the self-oscillation signal does not obscure the aerial signal. Two types of super-regenerative detectors are described in the following paragraphs. The first type employs a separate valve circuit called a *quench oscillator* for controlling the oscillation in the detector circuit. The second type is arranged to control its oscillation by means of a *squegging circuit.*

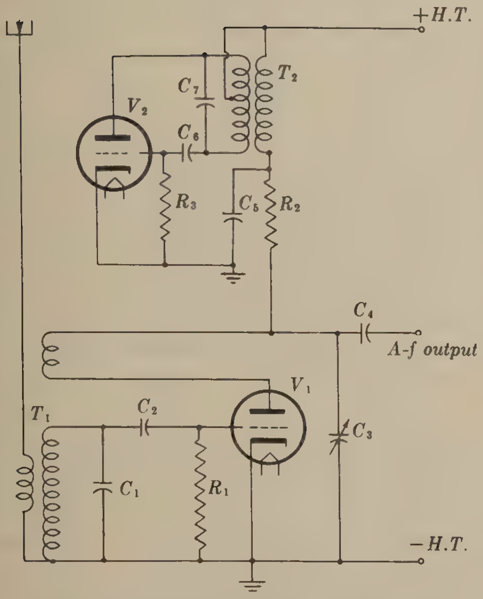


FIG. 9-13 Super-regenerative receiver with a quench oscillator.

A. Quench-Oscillator Receiver

A super-regenerative receiver which contains a quench oscillator is shown in Fig. 9-13. The circuit containing V_1 is a regenerative grid-leak detector similar to that shown in Fig. 9-7. The regeneration control is adjusted so that when V_2 is not operating and a steady value of H.T. is applied, the circuit of V_1 is on the point of oscillating. If the H.T. supply voltage is then increased, the regeneration increases and the circuit breaks into oscillation. If the H.T. supply voltage is reduced below the steady value, the oscillations are damped out.

The quench oscillator V_2 causes the supply voltage of V_1 to rise above and fall below the value required to cause oscillation. The circuit is a series-feed Hartley oscillator. It is made to operate at a frequency considerably above the audio range but far below the aerial signal frequency to which the detector is tuned. During the positive half cycle of oscillation in V_2 the anode voltage of V_1 is driven above its critical value by coupling through T_2. This permits oscillation to build up in the circuit of V_1 at the frequency set by T_1 and C_1. During the negative half cycle of oscillation in V_2 the anode potential of V_1 falls below the critical value and the oscillations in its grid circuit are damped out. Even when no signal is received in the aerial, oscillations are permitted to build up and die out in V_1 at the frequency of the quench oscillator. These oscillations, when detected, produce an output voltage through C_4 at the frequency of the quench oscillator. This frequency is maintained above the audio range to prevent an audio signal from being developed. The effect of the quench oscillator on the regenerative detector is illustrated by the waveforms of Fig. 9-14. The quench oscillator output voltage, oscillations in V_1 and output through C_4 when no aerial signal is received are shown in Fig. 9-10 (a), (b) and (c) respectively.

The effect of receiving an aerial signal modulated at an audio frequency is shown by the remaining waveforms of Fig. 9-14. In the absence of super-regeneration the a-f voltage output of V_1 follows the variations in the modulation envelope as shown in (d) and (e). With super-regeneration, as in (f), oscillations in the grid circuit of V_1 build up to a much greater amplitude during each positive half cycle of V_2. The mean amplitude of the oscillations in V_1 produced during one cycle of the quench oscillator is proportional to the instantaneous amplitude of the received signal, Fig. 9-14 (b). This gives rise to a much larger component of a-f current from the detector, Fig. 9-14 (g), than would be obtained without super-regeneration. While a super-regenerative receiver has a high degree of sensitivity, its selectivity is no better than an ordinary regenerative receiver.

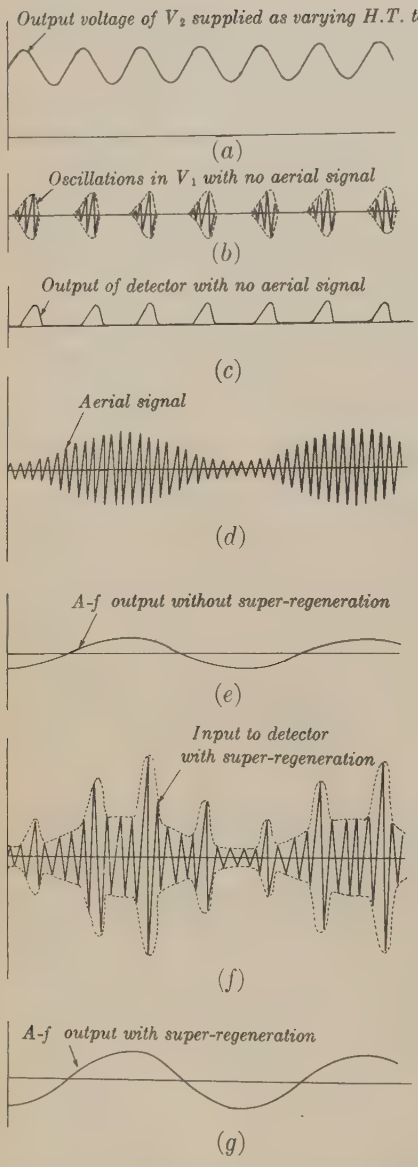


Fig.9-14 Voltage waveforms illustrating the operation of the quench oscillator and regenerative detector.

B. Self-Quenching Receiver

A simple super-regenerative detector of the self-quenching or squegging type is shown in Fig. 9-15. In this case, quenching is not performed by a separate oscillator but by intermittent blocking in the grid circuit. Sufficient feedback is provided by the tickler coil to make the circuit self-oscillating. In the absence of an aerial signal the circuit bursts into oscillation at the frequency to which the grid circuit is tuned. As the oscillation builds up rapidly, automatic bias is developed by C_2R_1 in the grid circuit. The grid is quickly driven so negative that the anode current is cut off. Oscillations cease and the circuit remains inoperative until C_2 discharges through R_1 sufficiently to permit anode current to flow again. Another burst of oscillation is then developed. The circuit therefore continues to produce bursts of oscillation separated by periods of inactivity. This action is termed *squegging*. The squegging frequency is determined by the time constant of C_2 and R_1 and is set above the audio range. This ensures that no a-f tone is produced by the detector when no signal is received.

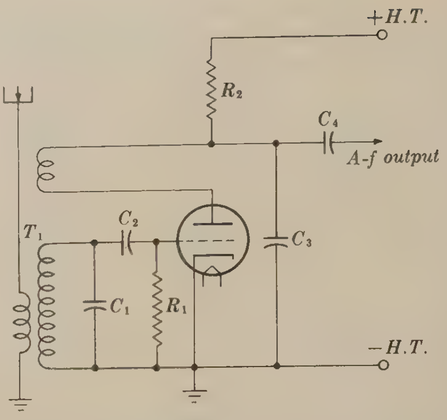


Fig.9-15 Self-quenching super-regenerative receiver.

When a signal is received in the aerial at the frequency to which the grid circuit is tuned, the circuit begins to oscillate earlier and oscillates with greater amplitude during each burst. The mean amplitude of oscillations during each burst is proportional to the instantaneous amplitude of the signal received. These bursts of oscillation when detected produce an a-f component of current which follows the modulation envelope of the signal. The amplitude of this a-f component of current is increased by the super-regenerative action of the circuit. Thus the sensitivity of the circuit is increased. The selectivity is no better than that of a regenerative receiver. In addition, a continuous background noise is present in the receiver when no signal is received.

A type of squegger used to receive c-w signals is arranged with the squegging frequency in the audio range. When a signal is received, the squegging frequency rises since the circuit bursts into oscillation more readily than when no signal is received. Consequently the pitch of the a-f tone rises for the duration of the dot or dash in the signal. Such a circuit is known as a *howling squegger*.

9.9 The superheterodyne principle. The two chief essentials of a good communications receiver are high selectivity and high sensitivity.

The selectivity of a receiver depends mainly on the number of tuned circuits and the selectivity of each. The extent to which a circuit can differentiate between signals depends not so much on their actual difference as on their percentage difference in frequency. It is easier to design a circuit to separate signals of 99 and 100 Kc/s than it is to separate signals of 999 and 1000 Kc/s. In each case the difference is 1 Kc/s. In the former the percentage difference is 1% while in the latter the percentage difference is only 0.1%. Good selectivity is thus more easily attained at low frequencies.

The sensitivity of a receiver depends upon the number of amplifiers used and on the gain of each amplifier. Interelectrode and stray capacitances are effectively in parallel with the load of an r-f amplifier and thus limit its impedance to a relatively low value at high frequencies. Since the gain of an amplifier depends to a great extent upon the a-c impedance of its load, it is difficult to design amplifiers to produce high gain at high radio frequencies. An r-f amplifier which can provide a gain of 200 at 300 Kc/s may give a gain of only 10 at 30 Mc/s.

The superheterodyne receiver was developed in an attempt to attain good selectivity and good sensitivity at high frequencies. It makes use of the heterodyne principle described previously to convert the signal from one of high frequency where the sensitivity and selectivity of the receiver are poor to one of lower frequency where the sensitivity and selectivity are good. The advantages of amplification before the audio detector stage are thus preserved while some of its shortcomings are eliminated.

A block diagram of the superheterodyne receiver is shown in Fig. 9-16. The aerial, tuner, second detector, b.f.o., a-f voltage amplifier, a-f power amplifier and speaker serve precisely the same functions as in the t-r-f receiver previously described.

The r-f amplifier, besides providing some amplification, serves to reduce certain types of interference peculiar to the superheterodyne receiver. Interference is discussed in a later section. The r-f amplifier may consist of one or more tuned stages which are ganged so that they may be tuned to the same frequency by one control. Pentode valves are generally used and the stages are well shielded and carefully laid out to prevent any possible parasitic oscillations.

The *local oscillator* is a heterodyne unit which produces a c-w signal of constant amplitude. This signal is made to differ in frequency from the aerial signal by a fixed amount, usually about 460 Kc/s. The local oscillator output is combined with the r-f signal in a *mixer circuit.* The combined signal then consists of a wave which is modulated at 460 Kc/s in addition to the a-f modulation pro-

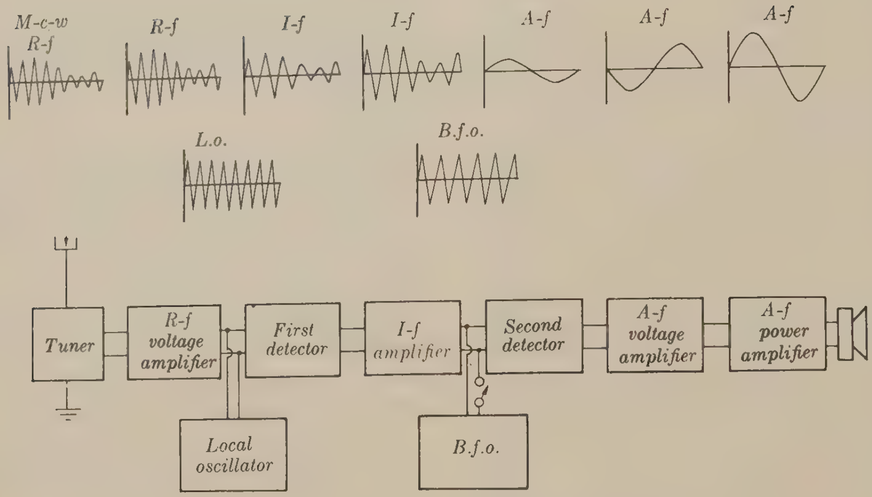


FIG.9-16 Block diagram of a superheterodyne receiver.

duced at the transmitter. The new signal is detected by the first detector to produce a carrier of frequency 460 Kc/s modulated at a-f. This signal is intermediate in frequency between the r-f carrier wave from the transmitter and the a-f output of the speaker. It is called the *intermediate frequency* or i-f. The tuning capacitor of the local oscillator is ganged with the tuning capacitor of the r-f amplifier so that they are always in tune to frequencies which are 460 Kc/s apart. The local oscillator is usually tuned to a higher frequency than the r-f amplifier. The following table shows how the local oscillator frequency is always kept 460 Kc/s above the r-f signal frequency so that the i-f remains constant.

R-F SIGNAL	L.O. SIGNAL	I-F SIGNAL
500 Kc/s	960 Kc/s	460 Kc/s
600	1060	460
700	1160	460
800	1260	460
900	1360	460
1000	1460	460

Careful alignment of the tuned circuits in the r-f amplifier and local oscillator stages is required in order to maintain the i-f constant over a wide band of frequencies.

The superheterodyne receiver derives its name from the fact that the intermediate frequency is in the supersonic range. The receiver was originally called the *supersonic heterodyne* receiver but is now called the *superheterodyne receiver* or simply the *superhet.*

A more detailed picture of the operation of the receiver is provided by the waveforms of Fig. 9-17. An r-f signal at 900 Kc/s is received by a superhet having an i-f of 460 Kc/s and an a-f signal of 2 Kc/s is produced in the output. It should be noted that no attempt has been made to maintain the same scale of amplitude throughout the diagrams. The voltage gain of the receiver may be as high as a million but it is impracticable to show the initial and final voltages in such a proportion. For similar reasons, no attempt is made to maintain the same time units throughout the diagrams for to do this would require that 450 complete cycles in diagrams (a), (b) and (c) be placed in the same horizontal

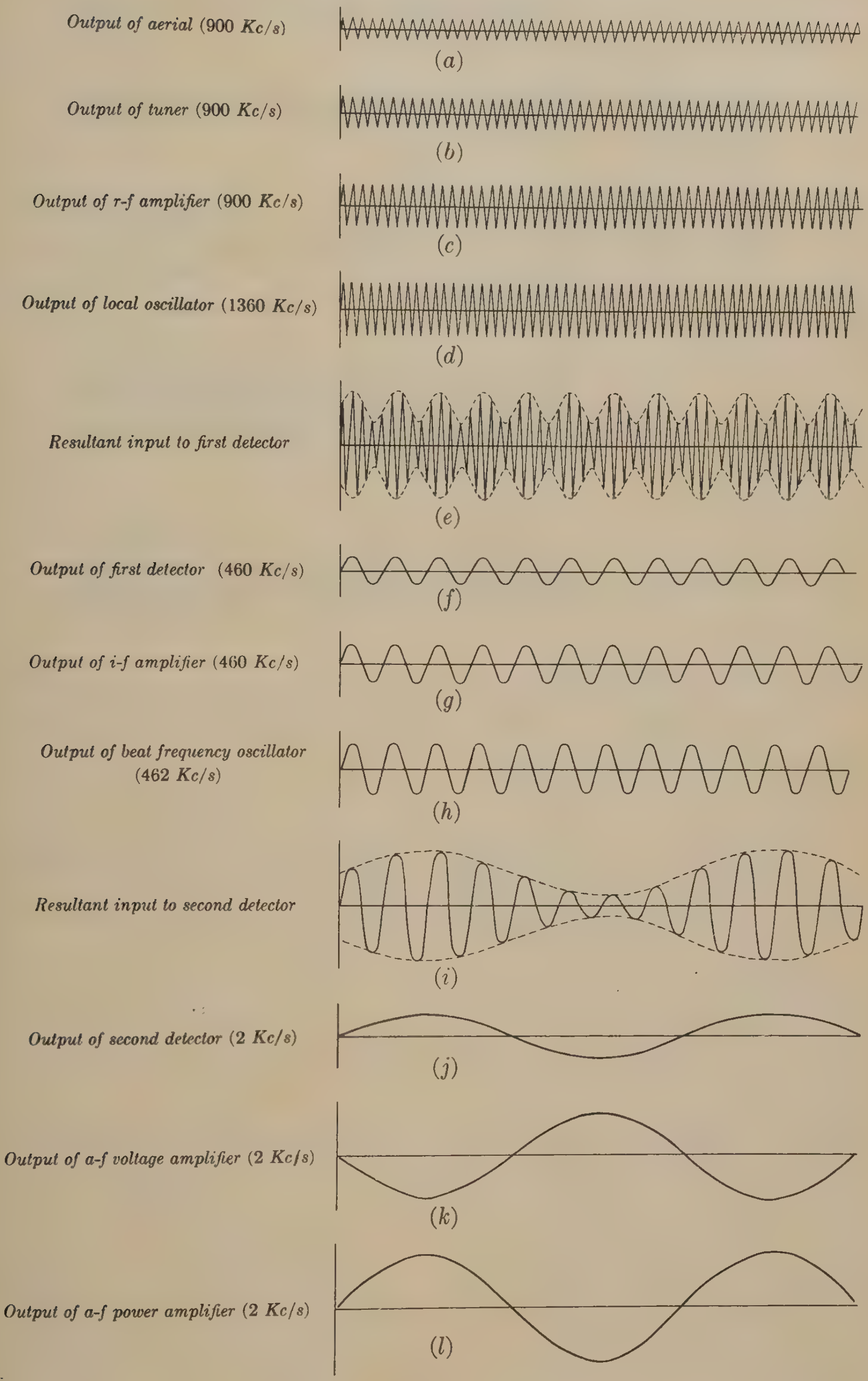

Output of aerial (900 Kc/s)
(a)
Output of tuner (900 Kc/s)
(b)
Output of r-f amplifier (900 Kc/s)
(c)
Output of local oscillator (1360 Kc/s)
(d)
Resultant input to first detector
(e)
Output of first detector (460 Kc/s)
(f)
Output of i-f amplifier (460 Kc/s)
(g)
Output of beat frequency oscillator (462 Kc/s)
(h)
Resultant input to second detector
(i)
Output of second detector (2 Kc/s)
(j)
Output of a-f voltage amplifier (2 Kc/s)
(k)
Output of a-f power amplifier (2 Kc/s)
(l)

FIG.9-17 Input and output voltages for the superheterodyne receiver stages with c-w signals.

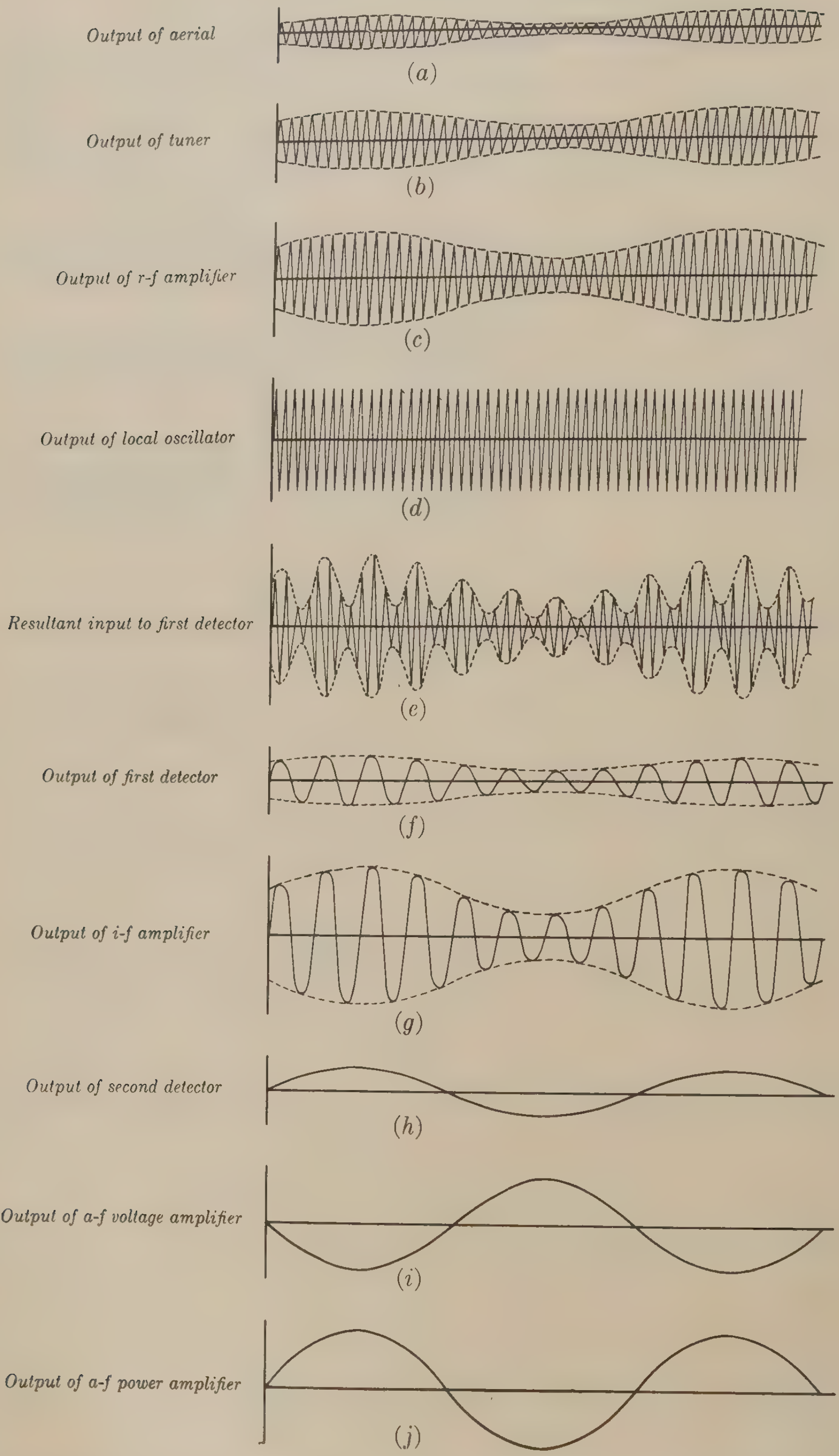

Output of aerial
(a)
Output of tuner
(b)
Output of r-f amplifier
(c)
Output of local oscillator
(d)
Resultant input to first detector
(e)
Output of first detector
(f)
Output of i-f amplifier
(g)
Output of second detector
(h)
Output of a-f voltage amplifier
(i)
Output of a-f power amplifier
(j)

FIG. 9-18 Input and output voltages for the superheterodyne receiver stages with m-c-w signals.

space as 230 cycles in diagrams (f), (g) and (h) and one cycle in diagrams (j), (k) and (l). Nevertheless, it is instructive to consider the operation of the set when receiving a signal at a specific frequency.

Fig. 9-17 (a), (b) and (c) show the voltage waveforms of a c-w signal at 900 Kc/s in the aerial circuit, the tuner and the r-f amplifier respectively. The output voltage of the local oscillator is shown in diagram (d). This signal is a c-w at 1360 Kc/s which is normally larger in amplitude than the r-f signal. These two signals are mixed together in the input circuit of the first detector. The resulting signal, shown in (e) is a continuous wave, modulated at 460 Kc/s which is the *difference* between the frequencies of the aerial and l.o. signals. The output voltage of the first detector is shown in (f). This is the intermediate-frequency signal, a c-w at a frequency of 460 Kc/s. The i-f signal is amplified as shown in (g) and mixed with the b.f.o. output, shown in (h), at 462 Kc/s. The resultant waveform, shown in (i), is a c-w which is modulated at 2 Kc/s. The second detector produces the modulation envelope of this waveform as shown in (j). It is the audio-frequency signal at 2 Kc/s. The a-f output is amplified as shown in (k) and (l) and produces the audio tone in the headphones or loudspeaker.

The waveforms of Fig. 9-18 are similar to those of Fig. 9-17 but show the operation of the set during the reception of an m-c-w signal. In this case a more complex waveform is developed in the input circuit of the first detector. The mixed signal is modulated at both the intermediate and the audio frequencies. The i-f transformer is tuned to the intermediate frequency to retain the i-f signal.

The superhet has certain definite advantages over the t-r-f receiver. The design of the i-f amplifier (Fig. 9-19) is simplified because it operates always at the same frequency regardless of the frequency of the r-f signal. For this reason the superhet is commonly used even at lower frequencies where other types of receivers give as good selectivity and sensitivity. Also, the b.f.o. may be preset to a fixed frequency which differs from the i-f by any desired audio frequency. Thereafter, tuning of the b.f.o. is not required unless it is desired to change the pitch of the audio tone.

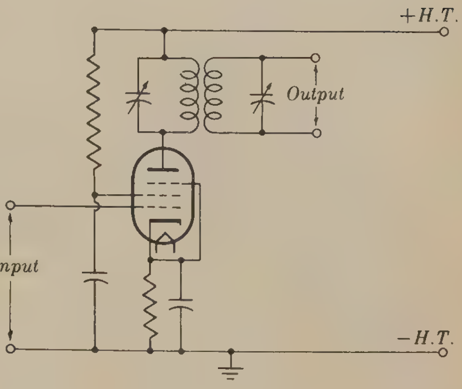


FIG.9-19 I-f amplifier circuit.

Of greater importance still is the high degree of selectivity attainable in the i-f amplifier circuits. This is achieved by transformer coupling between stages. Both primary and secondary circuits of the i-f transformers are tuned to the intermediate frequency. Hence a high gain is provided at the i-f while very little amplification is provided at neighbouring frequencies which are likely to cause interference in t-r-f receivers. The tuning of the i-f transformers is effected by varying the capacitance of the capacitors as shown in Fig. 9-19 or by altering the permeability of the cores in the coils.

The band-pass of the receiver is also determined to a large extent by the i-f transformers. The coupling between the primary and secondary windings is fixed for optimum band-pass. It may be made extremely narrow for the reception of c-w signals under conditions of extreme interference or very wide to permit the passing of sideband frequencies in the reception of high fidelity r-t signals. Some receivers are provided with several i-f circuits which may be switched into use for providing different band-pass characteristics. As in the case of the r-f amplifier, careful shielding and layout of component parts is necessary to prevent undesired oscillations from taking place.

9.10 Frequency converter circuits. As previously stated the principle of the superheterodyne receiver is to reduce the frequency of the incoming carrier from r-f to i-f. The stage which performs this function is known as the *frequency converter* stage. It consists of the local oscillator and the *first detector* or *mixer*.

The general principle of the action is as follows. The input circuit of the first detector is tuned to the frequency of the incoming signal. The local oscillator is tuned to a frequency which differs from the r-f signal by the intermediate frequency. The two signals are mixed together in the first detector and the result of this heterodyning is a wave modulated at the i-f as shown in Fig. 9-17 (e) and Fig. 9-18 (e). The non-linear characteristic of the first detector causes amplitude distortion. As a result, the anode current of the detector contains components at the sum and difference frequencies as well as at the two input frequencies and harmonics of all of these. Since the difference frequency is the frequency required, the anode load of the detector is tuned to this frequency which is the i-f. Thus a large output at the i-f is obtained while all other frequency components are effectively suppressed.

If the incoming signal is modulated (Fig. 9-18) each of the sideband frequencies is reduced along with the carrier frequency. Therefore the i-f has the same sidebands as the r-f and hence carries the same modulation.

In the design of frequency converters it is necessary:

(i) to maintain a constant frequency difference between oscillator and input circuits across the entire band
(ii) to minimize variations in oscillator frequency with variations in supply voltage, bias, valve electrode spacings, etc., any of which may cause frequency drift
(iii) to maintain a constant output voltage from the oscillator across the entire band
(iv) to keep radiation from the oscillator to a minimum in order to prevent interference in nearby receivers
(v) to isolate the input and oscillator circuits to avoid the effect of locking or pulling whereby the oscillator becomes synchronized with the incoming signals
(vi) to maintain the output as free as possible from harmonics which might cause interference.

Three main types of frequency converter systems have been devised using special methods and valves.

A. Oscillator-Detector Type

In this type of circuit, a separate valve oscillator feeds into a conventional detector. The oscillator may be of any type although tuned-grid oscillators and Hartley oscillators are most commonly used. The detector is usually a pentode acting as an anode-bend detector; at higher frequencies diode and crystal detectors are used. The oscillator may be coupled to the detector in any convenient way, provided that the voltage introduced is approximately constant. Mutual inductive coupling in the cathode circuit or capacitive coupling to the control or screen grids of the detector are the most common. In the oscillator-detector circuit of Fig. 9-20, a Hartley oscillator V_2 is capacitively coupled by C_3 to the screen grid of the detector V_1. The r-f signal is applied to the control grid of

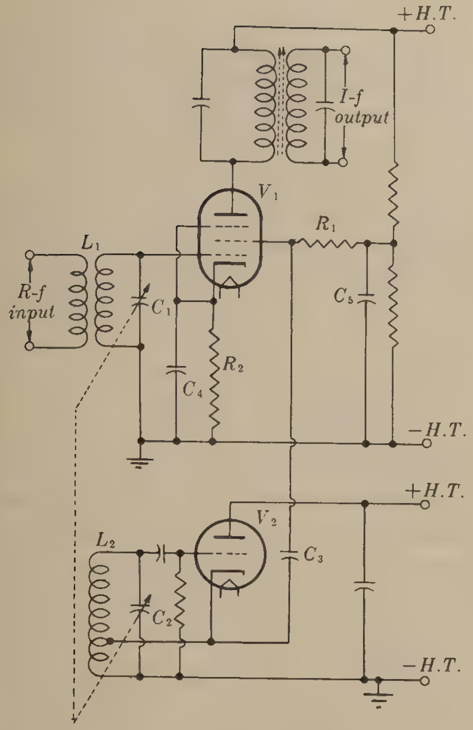


FIG.9-20 Oscillator-detector circuit.

V_1 and mixing occurs in this valve. The anode circuit is tuned to the i-f and is transformer coupled to the next stage. Capacitors C_1 and C_2 are ganged and designed so that the frequencies to which L_1C_1 and L_2C_2 are respectively tuned, always differ by the i-f over the tuning range. R_1 is present to prevent the oscillator signal from being grounded by C_5. R_2 and C_4 provide bias for V_1 causing it to operate as an anode-bend detector.

This type of circuit is quite sensitive to coupling between the input circuit and the oscillator circuit. At high frequencies where there is only a small percentage difference between the two frequencies, a strong signal may cause the oscillator to synchronize with the r-f so that the proper i-f is not produced. Also the possibility of radiation from the oscillator is somewhat greater in this type than in the other types of converters.

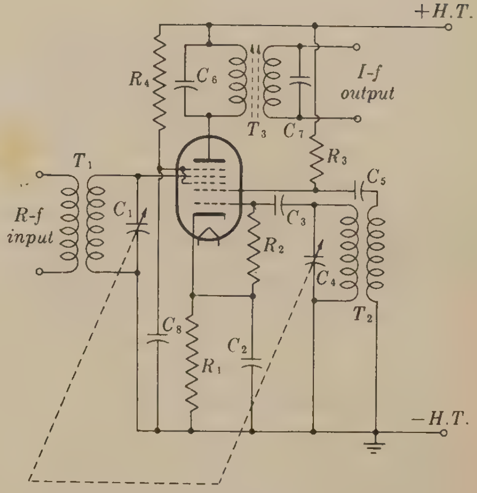


FIG.9-21 Pentagrid converter, triode-tetrode type.

B. Converter-Valve Type

The functions of oscillator and detector are performed by the same valve in the circuit shown in Fig. 9-21. The valve is specially designed for this purpose and contains five grids. It is called a *pentagrid* valve. The cathode and the first two grids perform the functions of cathode, grid and anode in a tuned-grid oscillator circuit. The frequency is determined by C_4 and the primary winding of T_2. The secondary winding of T_2 provides the feedback to maintain oscillation. Automatic bias is provided by C_3 and R_2. R_3 and C_5 are the anode load resistor and blocking capacitor of the oscillator. The electron stream through the second grid varies at the frequency of this tuned grid or local oscillator.

The r-f signal from the preceding circuit is applied through T_1 to the fourth grid of the valve. This causes the electron stream to vary at the r-f. In effect then, the electron stream reaching the anode is modulated at the i-f by the heterodyning action of the two circuits. The valve is biased so that detection takes place and the i-f component of

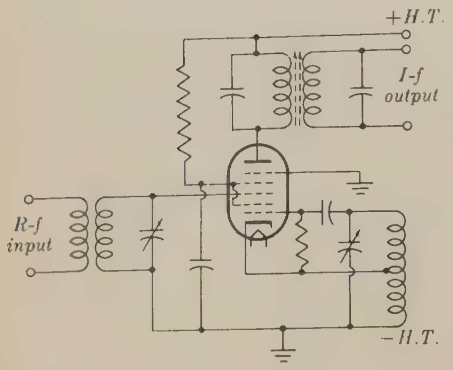


FIG.9-22 Pentagrid converter, triode-pentode type.

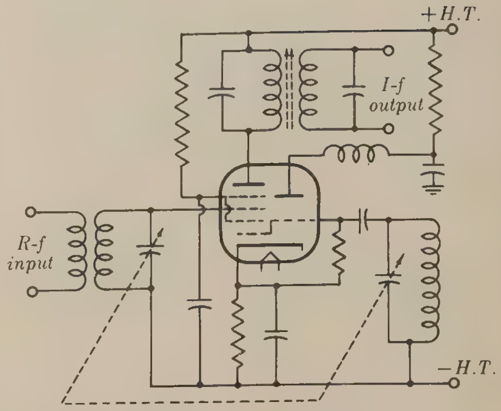


FIG.9-23 Oscillator-mixer with a triode-hexode.

current develops the intermediate-frequency voltage across the primary winding of T_3. The third grid serves as a shield between the oscillator and detector circuits. The fifth grid serves as a screen between the signal grid and the anode, giving to the valve the characteristics of a tetrode. R_4 and C_8 maintain the screen grids at a steady d-c potential. This connection of the pentagrid valve is referred to as the *triode-tetrode* connection.

A somewhat different connection of the pentagrid converter is shown in the circuit of Fig. 9-22. The cathode, first and second grids make up the oscillator section as before. In addition the second grid acts as the shield between the oscillator and detector circuits. This requires a modified type of oscillator in which the anode operates at r-f ground potential. In this circuit a modified Hartley oscillator is used. The third grid is used as the r-f signal grid while the other two grids serve as screen and suppressor. The detector therefore has the characteristics of a pentode. This connection of the pentagrid valve is known as the *triode-pentode* connection. The sensitivity of this circuit is good at low and medium radio frequencies. At high frequencies however, the sensitivity falls off and coupling between the two sections causes difficulty in stabilizing the oscillator output frequency and voltage. For these reasons, the converter valve circuit is confined to use at medium radio frequencies and the following type is used at high frequencies.

C. Oscillator-Mixer Type

The oscillator-mixer type of frequency converter employs a multi-element valve called a *triode-hexode.* It is shown in the circuit of Fig. 9-23. The triode oscillator section is in the same envelope as the detector section but is quite separate from it. Coupling between the two sections due to a common electron stream is not present in this circuit. The oscillator is of the conventional tuned-grid type. The output from the grid of the triode section is applied to the first grid of the hexode. The electron stream is therefore varied at the two frequencies producing a modulated signal. Grids two and four serve as screens between the two signal grids in front of the anode. The hexode operates as a detector as well as a mixer and the i-f component of current develops the i-f signal in the anode circuit.

The sensitivity of the triode-hexode is slightly less than that of the oscillator-mixer type, but the circuit can be used satisfactorily on higher frequency bands. The noise level is low due to the reduced coupling between the oscillator and the signal grids. The circuit is very useful in receivers lacking an r-f stage as there is little radiation from the oscillator through the aerial circuit.

Generally the local oscillator is designed to operate at a higher frequency than the signal frequency. The reason for this may be seen by considering a broadcast receiver with an i-f of 460 Kc/s, operating from 550 Kc/s to 1700 Kc/s. If the oscillator operates above the signal frequency, it must tune from 1010 Kc/s to 2160 Kc/s. If it is operated below the signal frequency it would have to tune from 90 Kc/s to 1240 Kc/s. In the former case, the upper frequency limit of the local oscillator is only twice the lower frequency limit. In the latter case the upper limit is about 14 times the lower limit. Since the upper frequency limit of the broadcast band is only about three times the lower limit, it is easier to design an oscillator to track with an r-f amplifier when the l.o. operates above the signal frequency.

9.11 Alignment of the superheterodyne receiver. For optimum operation of the superheterodyne, the i-f amplifiers must be tuned accurately to the i-f. The local oscillator must be tracked with the first detector so that all the way across the band the two differ in frequency by the i-f. Finally, the r-f amplifiers and first detector must be tracked so that they are always tuned to the same frequency. The process of making the above adjustments is called the *alignment* of the receiver.

The procedure may be divided into two sections:

(i) the i-f alignment
(ii) the r-f and oscillator alignment.

The equipment necessary is a calibrated signal generator and some means of measuring the output of the receiver. A tuning indicator on the receiver will serve the latter purpose or, if a modulated signal generator is used, an output meter, oscilloscope or vacuum-tube voltmeter across the load of the a-f power amplifier. If these latter are not available, the output may be roughly estimated by ear. While the alignment procedure outlined in the following paragraphs is applicable to most broadcast receivers, communications receivers should be aligned in accordance with the manufacturers' instructions.

A. I-F Alignment

The ground lead of the signal generator is connected to the chassis and the r-f off-ground lead is connected to the grid of the last i-f amplifier. The signal generator is adjusted to the desired i-f. The tuning of the i-f coupling transformer between this amplifier and the second detector is now adjusted for maximum output. The secondary is adjusted first and then the primary. Readjustment of both circuits should be made after the preliminary adjustment is completed. The off-ground lead of the generator is now removed from the last i-f amplifier and connected to the grid of the preceding i-f amplifier. The i-f coupling transformer between the last two i-f amplifiers is then tuned for maximum output as before. In this manner all the i-f transformers are aligned working back from the second detector to the first. As the amplifiers are brought into tune it is necessary to reduce the output of the signal generator to prevent overloading and inaccurate readings. If the tuning indicator is used as an output meter, a.v.c. must be switched on; if an audio output method is used the a.v.c. should be turned off. In each case the b.f.o. should be turned off.

B. R-F Alignment

The main tuning control should be set at the high-frequency end of the dial for the range in use. The signal generator

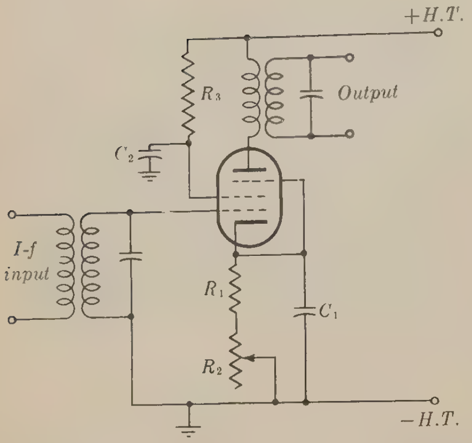


FIG.9-24 I-f amplifier with manual volume control in the cathode circuit.

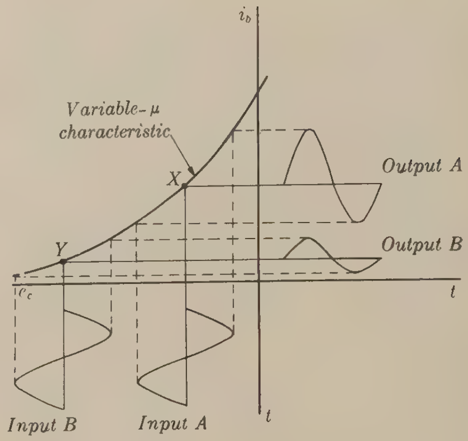


FIG.9-25 An e_c—i_b curve illustrating variable-mu operation.

is connected to the aerial input through a dummy aerial and adjusted to the correct frequency. The oscillator tuning is now adjusted by means of the trimmer capacitors for maximum output. Next, the trimmer capacitors on the first detector and r-f amplifiers are adjusted in the order mentioned for maximum output. The tuning dial and signal generator are reset to the bottom end of the dial and the procedure repeated. If the coils have variable cores or padder capacitors, the tuning adjustments at the low end of the band should be made by means of these rather than by the trimmer capacitors.

The above procedure must be repeated several times until the alignment of all circuits is the best that can be obtained throughout the whole tuning range. It is seldom perfect over the entire band. However, the variation in gain caused by slight mistuning at any point in the band is usually negligible. If the major portion of the reception is to be done at some small part of the band it may be advantageous to align the receiver accurately at that part of the band and permit slight mistuning in regions that are seldom used.

9.12 Manual volume control. The power output of a receiver may be controlled either manually or automatically. Manual control is usually effected by changing the gain of one or more stages of the i-f or r-f amplifiers. Several variations of this method are described in the following paragraphs.

An i-f amplifier with manual volume control is shown in Fig. 9-24. The volume control is a rheostat in the cathode circuit of the valve. It is used to change the bias on the stage. The valve is a variable-mu pentode. When the resistance of R_2 is a minimum, the bias on the valve is small and the operating point is well up on the mutual characteristic curve. This is shown as point X in Fig. 9-25. At this point the mutual conductance of the valve is large. A small change in grid voltage causes a relatively large change in anode current and so the gain of the stage is high. If the resistance of the cathode circuit is increased, the bias becomes greater. The operating point shifts to a lower part of the characteristic curve shown as point Y. In this case input B, which has the same amplitude of grid voltage variations as input A, produces a much smaller change of anode current. The gain of the stage is reduced correspond-

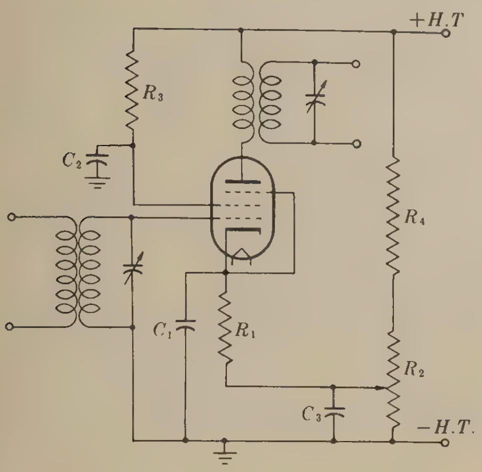


FIG.9-26 I-f amplifier with manual volume control in the cathode circuit.

ingly. The resistor R_1 in series with the gain control ensures that the bias is never reduced to zero. A volume control which operates in the r-f or i-f amplifier is usually labelled R-F GAIN.

A variation of the above method of gain control is shown in the circuit of Fig. 9-26. The lower end of the cathode circuit is returned to the volume control R_2 which is part of a voltage divider across the H.T. supply voltage. If the sliding contact is moved up, the positive potential on the cathode is increased. Since the grid is returned to ground for d-c, the bias is increased. This results in less amplification as explained for the previous circuit.

The volume control may also operate in the screen circuit of a pentode valve as shown in Fig. 9-27. If the sliding contact of the control is moved up, the screen potential is increased. This causes the slope of the mutual characteristic to increase and hence the amplification to increase. The effect of the gain control is illustrated in Fig. 9-28 in which the same input voltage is applied to the circuit with different values of screen voltage. When the screen potential is high the mutual characteristic curve, illustrated by *A*, is steep. The variations in anode current and hence the gain of the stage is high. When the screen potential is reduced the mutual characteristic curve becomes flatter as indicated by *B*. The gain of the stage is correspondingly reduced.

A similar type of control may be obtained by varying the anode potential on the valve as shown in Fig. 9-29. When the anode supply voltage is increased, the mutual conductance of the valve becomes greater and so the gain becomes larger.

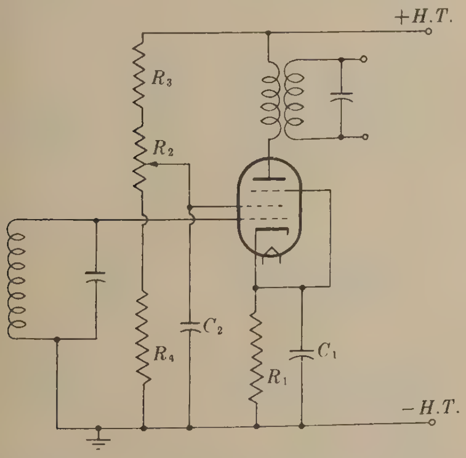


FIG.9-27 I-f amplifier with manual volume control in the screen circuit.

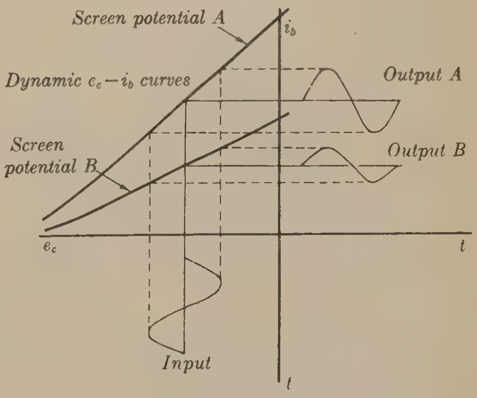


FIG.9-28 e_c—i_b curve illustrating screen voltage gain control.

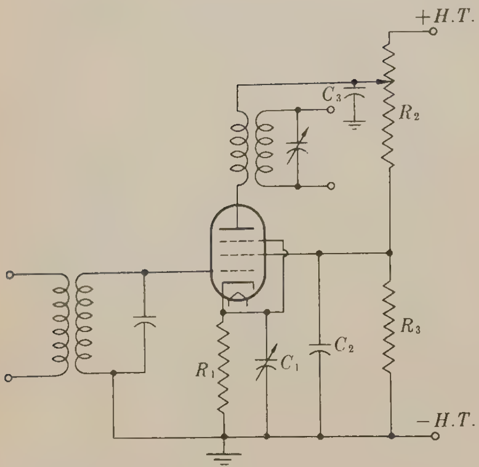


FIG.9-29 I-f amplifier with manual volume control in the anode circuit.

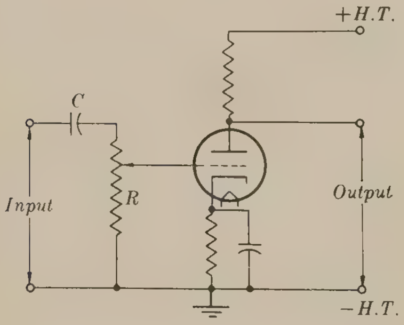


FIG.9-30 A-f amplifier with manual volume control in the grid circuit.

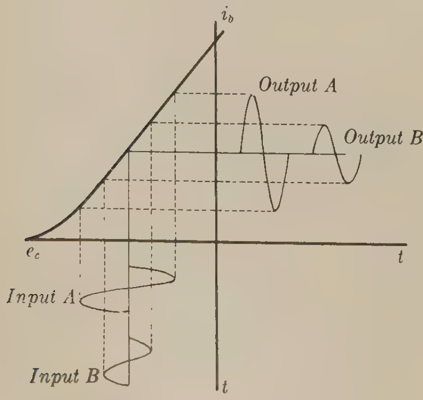
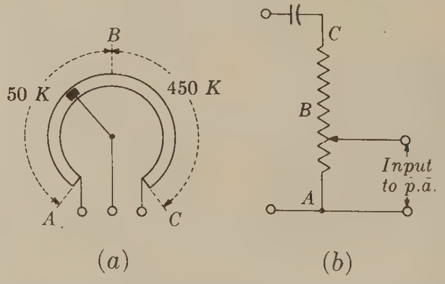


FIG.9-31 e_c—i_b curve illustrating manual volume control in the grid circuit.

Manual control of the volume may also be obtained by using a potentiometer to vary the magnitude of the input signal to one stage. This is commonly used in a-f amplifiers and the control is labelled A-F GAIN. A typical circuit is shown in Fig. 9-30. The a-f input signal is applied across *C* and *R* in series, most of the voltage appearing across *R*. When the sliding contact is at the top, all of the voltage across *R* is applied between grid and cathode of the valve and the volume is a maximum. This is illustrated by input *A* and output *A* in Fig. 9-31. As the sliding contact is moved down, the input voltage to the grid of the valve is reduced and so the output is also reduced. This is indicated by input and output *B* in Fig. 9-31.

Potentiometers used as volume controls are often constructed to have a *logarithmic taper* to compensate for the logarithmic response of the human ear to sound intensity. Consider the potentiometer shown in Fig. 9-32 in which the resistance in the first half of the rotation from *A* to *B* is 50 $K\Omega$ and in the second half of the rotation from *B* to *C* is 450 $K\Omega$. The resistance and hence the voltage input to the following amplifier is 10 times as great at setting *C* as it is at setting *B*. Therefore the power output from the amplifier at setting *C* is 100 times as great as at setting *B*. The logarithm of the power ratio is therefore 2, which means that the output at setting



(a) (b)

FIG.9-32 Potentiometer with a logarithmic taper (a) Plan view (b) Circuit diagram.

C sounds twice as loud to the human ear as the output at setting B. The apparent loudness therefore increases in proportion to the angle of rotation of the volume control.

9.13 Automatic volume control. Automatic volume control in a receiver is a method whereby the total amplification of the receiver is automatically reduced during the reception of a strong signal and increased during the reception of a weak signal. There are two reasons why a.v.c. is desirable:

(i) to obtain substantially the same a-f output from the receiver when the signal is fading

(ii) to prevent unpleasant blasts which might damage the receiver when tuning through stations of widely varying signal strengths.

Automatic volume control is achieved by applying an a.v.c. voltage to the grid of one or more variable-mu valves used as r-f or i-f amplifiers. The a.v.c. voltage is a negative d-c voltage which is proportional to the carrier strength of the received signal. It is applied as additional bias to reduce the gain of the amplifiers as the carrier strength increases.

A. *Types of A.V.C.*

1. SIMPLE A.V.C.

The simplest method of obtaining a.v.c. is to add a low-pass filter circuit consisting of a resistor and capacitor to the ordinary diode detector circuit as in Fig. 9-33. The negative d-c voltage obtained from this filter is then used for bias on the stages to be controlled, for example, the r-f amplifiers in the receiver.

The circuit is a conventional series diode detector in which R_1 is the load resistor and C_1 is the r-f by-pass capacitor. Across this load is developed an a-f voltage which is the output of the detector and a d-c voltage of polarity indicated in the diagram. If the input signal has a large amplitude this d-c voltage is large; if the signal is small this d-c voltage is small. As shown in Fig. 9-25 the stage gain of an amplifier using a variable-mu valve depends upon the bias. A strong signal produces a large negative bias which reduces the gain of the amplifier. A weak signal produces less bias causing the gain of the amplifier to increase. In this way the output of the receiver is held nearly constant over a wide range of signal strength.

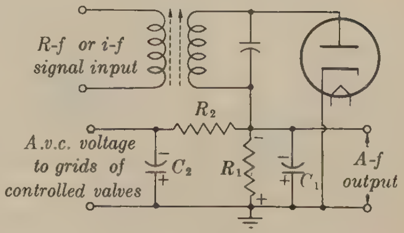


FIG.9-33 Diode detector with simple a.v.c.

If the low-pass filter R_2C_2 is omitted (Fig. 9-33) the a-f voltage across R_1 is fed back to the grids of the controlled amplifiers in addition to the d-c voltage. This means that the trough of the modulated signal is amplified more than the peak with a resultant smoothing of the modulation envelope. This effect is sometimes encountered in receivers when the a.v.c. filter capacitors have become open circuits. The time constant of R_2C_2 is usually of the order of one-half to one-twentieth of a second. This permits the a.v.c. to shift bias several times per second to compensate for fading. On the other hand, the time constant is long enough to prevent the bias from shifting with sufficient rapidity to remove even low-frequency modulation from the r-f signal.

2. DELAYED A.V.C.

A.v.c. circuits may be designed so that no a.v.c. bias is applied until the signal strength reaches a pre-determined value. These circuits are called *delayed a.v.c.*

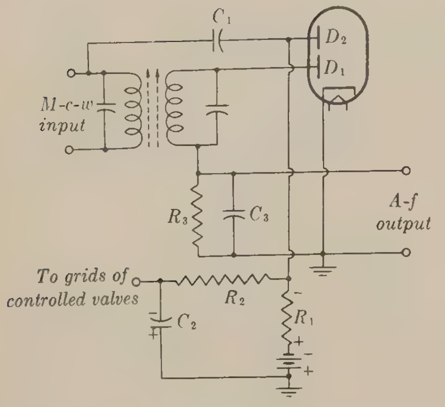


FIG. 9-34 Diode detector with delayed a.v.c.

(d.a.v.c.) circuits; the delay refers to voltage and not to time.

There are several advantages of d.a.v.c. over simple a.v.c. The sensitivity of the receiver is always greater for d.a.v.c. than for simple a.v.c. This is especially the case for small signals where d.a.v.c. approaches no-control conditions. By the use of a separate detector, d.a.v.c. may be arranged to provide a fairly large control voltage. Since it does not appreciably load the detector, distortion effects are very small. To reduce the loading effect further the signal for the d.a.v.c. detector may be taken from the primary of the i-f coupling transformer. A typical circuit is shown in Fig. 9-34 in which separate diode detectors are used to obtain the a-f signal and the d.a.v.c.voltage.

D_1 is the usual type of series-diode detector providing an a-f output across its load resistor R_3. D_2 is the d.a.v.c. detector. The negative delay voltage applied to the anode circuit of D_2 prevents the diode from conducting unless the peak input voltage exceeds the delay voltage. For weak signals D_2 does not conduct and thus no bias voltage is developed across R_1. For stronger signals, however, the circuit behaves as a parallel diode detector, developing a d-c voltage across R_1 with polarity as shown. This is the control bias and is fed back to the controlled amplifiers as in simple a.v.c. R_2C_2 is the low-pass filter which prevents a-f control.

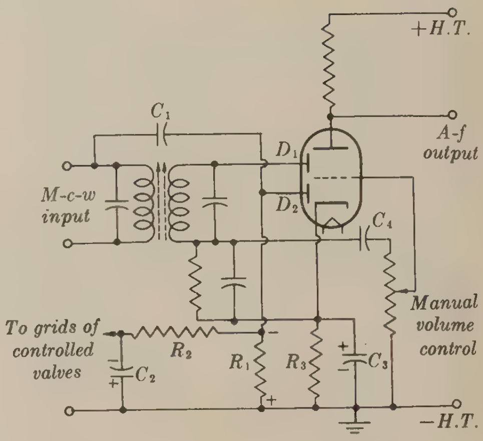


FIG. 9-35 Detector and a-f amplifier with delayed a.v.c.

Fig. 9-35 illustrates a practical method of producing d.a.v.c. in conjunction with the a-f detector and voltage amplifier. The output of the a-f detector D_1 is coupled through C_4 to the triode amplifier. The voltage in the primary of the i-f transformer is coupled through C_1 to D_2, the d.a.v.c. detector. The delay voltage is the bias voltage developed across the cathode resistor R_3 of the triode. If the potential of the anode of D_2 becomes more positive than the cathode potential, D_2 conducts and an a.v.c. output is developed across R_1.

3. AMPLIFIED A.V.C.

The a.v.c. control may be made more effective by producing a larger controlling voltage than can be obtained in the circuits discussed previously.

A separate a.v.c. channel may be used consisting of one or more amplifiers, in parallel with the signal channel. In Fig. 9-36, the signal is applied through C_1 to a single a.v.c. amplifier V_1. This amplifier has greater gain under all conditions than the signal amplifiers since there is no a.v.c. bias applied to it. Its output is coupled through C_2 to the anode of the

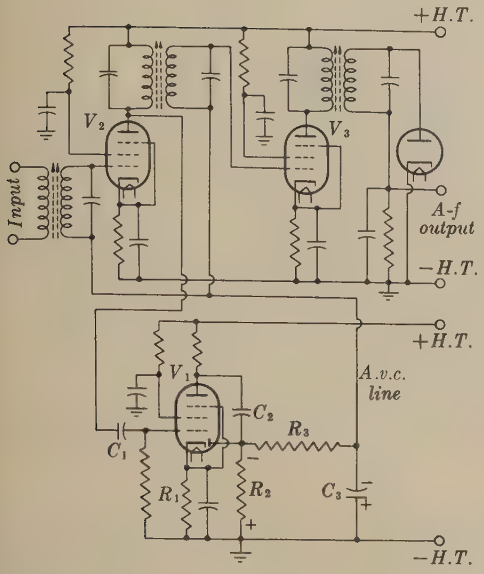


Fig.9-36 Amplified d.a.v.c. circuit using separate i-f channels.

diode section. The delay voltage is developed across the cathode resistor R_1. When the diode section conducts, the negative control voltage across R_2 is applied through the filter circuit R_3C_3 to the grids of V_2 and V_3. Since the control voltage is amplified, there is greater control of the i-f amplifiers V_2 and V_3. However, since the control voltage is taken from an earlier stage than in simple a.v.c. the selectivity is often inferior and the a.v.c. operates farther from the carrier frequency.

The control amplifier may be situated after the main i-f amplifier if a common i-f channel is used. Fig. 9-37 shows this control amplifier V_1 whose input is obtained from the last stage V_2 of the i-f amplifiers. The output from the control amplifier feeds the a.v.c. rectifier while the main amplifier feeds the signal detector.

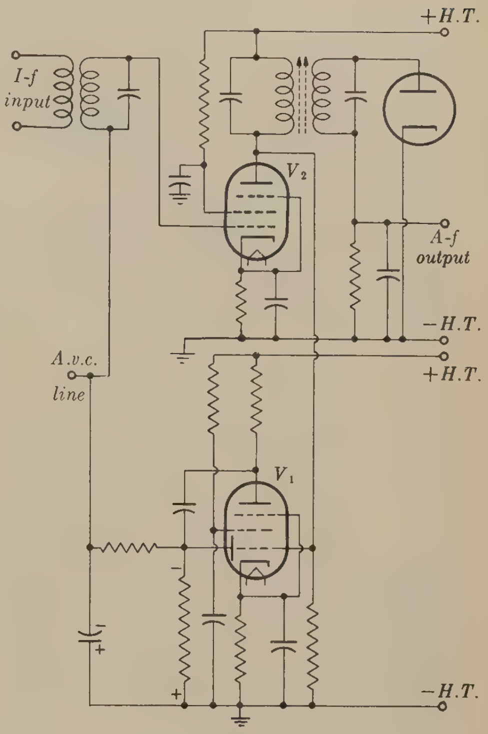


Fig.9-37 Amplified d.a.v.c. circuit using common i-f channel.

B. Methods of Feed

Either series or parallel feeding of the a.v.c. voltage to the controlled amplifiers may be used. These are illustrated respectively in Figs. 9-38 and 9-39.

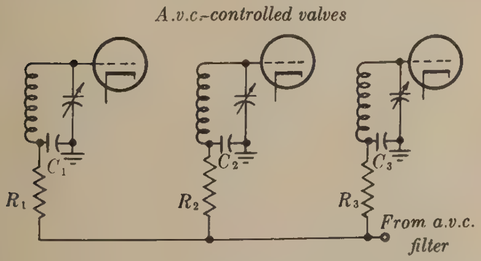


Fig.9-38 Series feed of a.v.c. voltage.

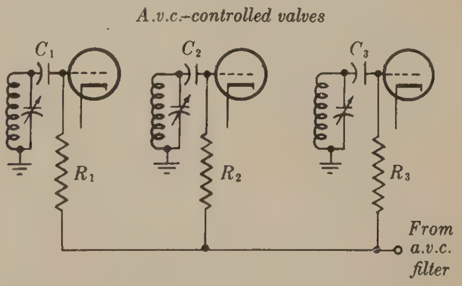


Fig.9-39 Parallel feed of a.v.c. voltage.

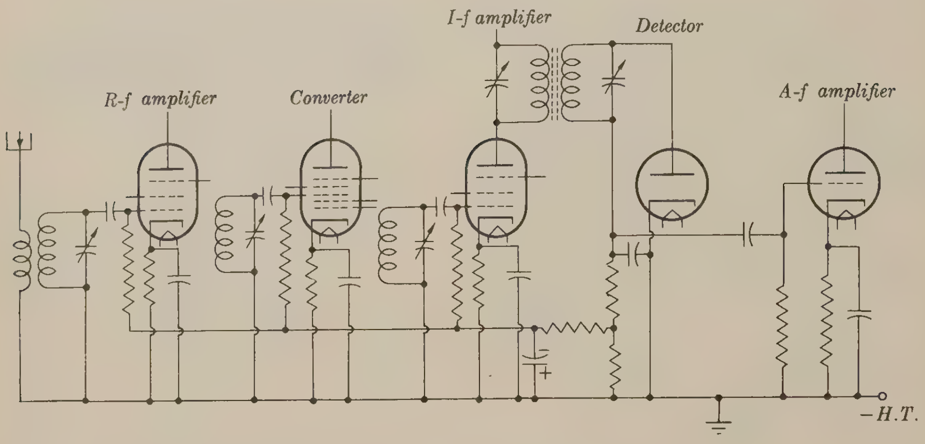


FIG.9-40 Simple a.v.c. circuit with parallel feed.

In the *series-feed* circuit (Fig. 9-38) the decoupling circuits R_1C_1, R_2C_2 and R_3C_3 prevent a-f modulation of the a.v.c. and prevent coupling between various stages using the a.v.c. bias. C_1, C_2 and C_3 complete the r-f tuned circuits and allow the rotor plates of the tuning capacitors to be grounded. In the *parallel-feed* circuit of Fig. 9-39, capacitors C_1, C_2 and C_3 are necessary to prevent shorting of the a.v.c. voltage to earth through the tuned circuit. Resistors R_1, R_2 and R_3 prevent grounding of the signal voltage through the a.v.c. filter.

C. Typical Circuits

Figs. 9-40 and 9-41 illustrate two circuits, the first using simple a.v.c. with parallel feed and the second using d.a.v.c. with series feed. Series feed is usually employed in the i-f stage since the damping effect produced by parallel feed is more objectionable in the i-f stages where selectivity is high.

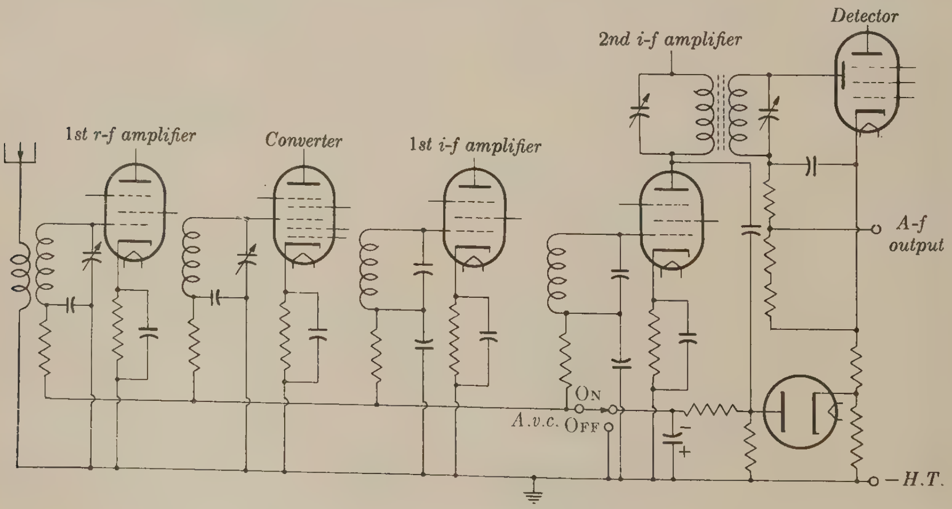


FIG.9-41 A d.a.v.c. circuit with series feed.

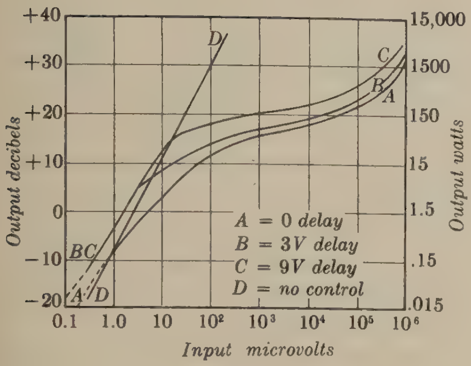


FIG.9-42 A.v.c. characteristics.

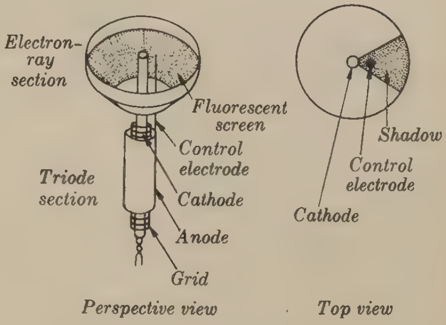


FIG.9-43 Electron-ray tuning-indicator valve.

D. A.V.C. Characteristics

Fig. 9-42 illustrates the relationship between the output and the input of a receiver under various types of a.v.c. control. The scale of power output in watts represents the output which would be obtained with the manual volume control at a maximum and 30% modulation, provided no over-loading occurred in the audio amplifier. It may be noted that when small signals are received, the output of the receiver is greater with a.v.c. than without. When large signals are received, the output of the receiver is, of course, reduced by the use of a.v.c. Hence the curves representing a.v.c. control cross over the curve representing no control near the low output end of the graph. This may be accounted for by the fact that a receiver with no a.v.c. must have a fairly large steady bias on the i-f and r-f amplifier stages in order to handle large signals as well as weak ones without distortion. Receivers using a.v.c. may use a smaller steady bias giving greater amplification of weak signals, since the extra bias necessary for strong signals is automatically supplied.

9.14 Tuning indicators. Tuning indicators are used to ensure that the receiver is accurately tuned to the signal frequency. They may also be used to give a measure of the relative strengths of received signals. There are two types of indicators in general use, the *electron-ray type* (*magic eye*) and the *output* or *S meter*. Fig. 9-43 shows a typical magic eye. Its construction and operation are described in detail in Chapter I. It consists of two parts, a triode amplifier and an electron-ray indicator, both using the same cathode. The anode or target of the indicator section is a fluorescent screen kept at a positive potential. Electrons emitted by the cathode are attracted to the target by the positive voltage. The fluorescent screen glows where it is struck by electrons. The control electrode, which is mounted between the cathode and target is connected to the anode of the triode section from which it receives its potential. When the control electrode is less positive than the target it repels electrons from that sector to produce a shadow on the target behind it. The size of this shadow increases as the potential difference between the target and control electrode increases.

The valve is operated by a negative a.v.c. voltage applied to its grid (Fig. 9-44). With no input voltage a current flows through the triode section. This current develops a p.d. across R_1 which reduces the potential of the control electrode below that of the target. This produces a shadow of approximately 90°

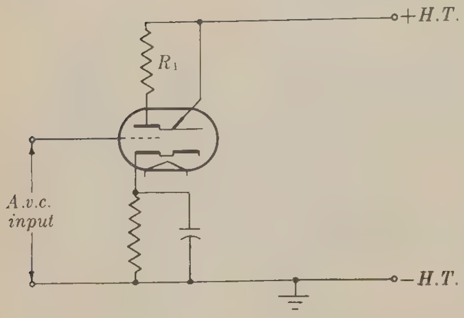


FIG.9-44 Tuning-indicator valve circuit.

on the screen. When a signal is correctly tuned in by the receiver, the a.v.c. voltage is at a maximum so that the current through the triode section is reduced and the anode voltage rises. The potential of the control electrode therefore rises and reduces the size of the shadow sector. This causes the eye to close. For accurate tuning over a large range of signal strengths, a potentiometer may be used in the grid circuit of the triode section. When a strong signal is being received, the potentiometer is set to reduce the a.v.c. input voltage to the tuning indicator. In this way the a.v.c. voltage is prevented from cutting off the current through the valve and the sensitivity of the tuning indicator is maintained.

A simple S meter circuit consists of a milliammeter in the anode circuit of an a.v.c. controlled valve. When the signal is properly tuned in, the bias is a maximum and the anode current drops. Unfortunately this type of meter reads backwards, a maximum signal causing a minimum reading on the meter.

Fig. 9-45 shows the circuit diagram of a forward-reading type of S meter. The positive terminal of the milliammeter is connected at A in the anode circuit of an amplifier to which a.v.c. is applied. The negative terminal of the meter is connected at B, a point of positive potential in the voltage divider across the H.T. supply. With no input signal, R_1 is adjusted so that the potential of point A is equal to that of point B. Thus the meter reading is zero. When a signal is tuned in, the a.v.c. causes the bias on the valve to increase and the anode current to decrease. This causes the potential of point A to rise and a current flows through the meter. The larger the signal, the greater this current becomes and therefore the greater the reading on the meter becomes.

9.15 Tone control. The tone control in an audio amplifier is designed to make the reproduced signal more intelligible or more pleasant to the ear. This may require the reduction in amplitude of some frequencies in the a-f signal and the boosting of others. In public address equipment, attenuation of the lower speech frequencies is provided to reduce the boom and echo which are so pronounced in a ship. Accentuating or boosting the higher speech frequencies is sometimes provided to emphasize the consonants and so improve the intelligibility of the speech. These relatively weak tones are otherwise lost because of the poor acoustic conditions existing in a ship. In receivers, the frequency response of a-f amplifiers is usually better than that required for the reception of speech. Because of this fact, the very high audio

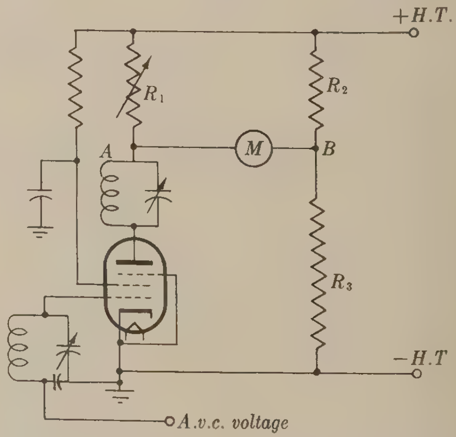


FIG.9-45 Circuit for a forward-reading S meter.

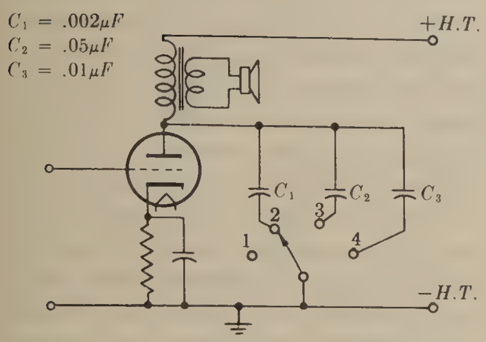


FIG. 9-46 A-f amplifier with capacitor tone control.

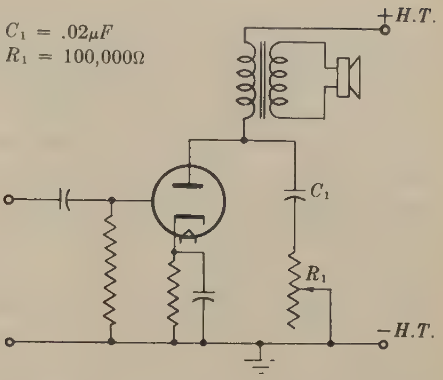


FIG. 9-47 A-f amplifier with capacitor-rheostat tone control.

frequencies may be attenuated to reduce static, while the lower ones may be eliminated to improve intelligibility.

High-frequency attenuation or *bass boosting* may be accomplished by the insertion of one or more capacitors across the output of an a-f amplifier. In Fig. 9-46, four different tone effects are obtained with the three capacitors and the four-position switch. With the switch in position 1, there is no tone compensation. In position 2, the reactance of the capacitor to low audio-frequency currents is so high in comparison with the load, that there is negligible by-passing of these components. At high audio frequencies, the reactance of the capacitor is small enough in comparison with the load that more of the current at these frequencies is by-passed to ground than is applied to the load. In position 3, still more of the high-frequency components are attenuated by the larger capacitor than in position 2. In position 4, an even larger proportion of currents in the upper frequency register are attenuated.

Another common tone control circuit which provides bass boosting is shown in Fig. 9-47. A rheostat R_1 and capacitor C_1 are connected in series from the anode of the power amplifier to ground. The rheostat permits a smooth variation in the tone of the audio output from the speaker. The entire resistance of R_1 is large in comparison with the reactance of C_1 to a-f currents. When R_1 is set to include all its resistance in the circuit, the total impedance of R_1 and C_1 is so large that the by-passing effect is almost negligible. Hence all the components of a-f current are passed to the speaker. This is the treble position of the tone control. When the resistance of R_1 is reduced to zero, C_1 provides a low-impedance path to ground for high-frequency components of current. The low-frequency sounds then predominate in the speaker and the bass boosting is a maximum.

If it is desired to boost the treble sounds from the speaker, the low-frequency components of the signal are attenuated. This is done by making use of a grid coupling capacitor whose reactance is large in comparison with the resistance of the grid-leak resistor at the frequencies to be attenuated. In that case a large percentage of the signal voltage appears across the coupling capacitor and is not applied to the grid of the valve. For example, with a grid-leak resistor of 0.5 megohms and a coupling capacitor of .002 microfarads, a-f components below 400 c/s are effectively attenuated.

9.16 Interference. The word interference is used in a broad sense to refer to the masking of the receiver output by undesired signals and also by electrical disturbances from sources other than radio equipment. Interference between signals is described in the following paragraphs and the subject of noise is discussed in a subsequent section.

Much of the interference between signals arises from the heterodyne process so that the superhet receiver is the worst offender in this respect. However, this type of receiver has so many other advantages that it still remains the most useful receiver developed up to the present.

A. Types of Interference

1. ADJACENT-CHANNEL INTERFERENCE.

Interference between transmitters on neighbouring frequency channels is by far the most common and therefore the most important type of interference between signals. The obvious remedy is to obtain a highly selective receiver. The superheterodyne by the use of a low i-f has largely overcome this difficulty.

2. 10 Kc/s WHISTLE.

Transmitting stations in the broadcast band are allocated carrier frequencies at 10 Kc/s intervals. If two signals on neighbouring frequencies reach the audio detector with sufficient power, their carrier waves will beat to produce a 10 Kc/s note. This is best remedied by improving the selectivity of the set or by introducing a filter to remove the objectionable 10 Kc/s note. In the case of c-w transmitters the frequency channels may be much closer together and the audio note of correspondingly lower pitch. The selectivity of communications receivers must be much greater than that of broadcast receivers.

3. CROSS MODULATION.

The modulation of strong signals may be superimposed on the modulation of the desired carrier because of distortion in an r-f amplifier so that both signals are heard in the output. In the case of superhet receivers, a strong station may cause the oscillator output to be modulated so that an unwanted signal is present in the i-f modulation in addition to the desired signal. The use of shielding in the local oscillator and i-f stages together with a pre-selector stage before the frequency converter are effective means of eliminating cross modulation.

4. TRANSMITTER HARMONICS.

Transmitters are prohibited by law from radiating energy above a certain level at harmonics of their carrier frequencies. However, strong local stations may radiate sufficient power at harmonic frequencies to be objectionable in the receiver. Harmonics of a transmitter frequency may also be generated in the receiver itself by r-f amplifiers or first detectors which introduce distortion. The result is the same as if the transmitter were generating the undesired signals. Here again the most effective remedy is to provide some selectivity before the first detector and to use Class A r-f amplifiers.

The types of interference between signals thus far considered may be found in all types of receivers. The following faults are peculiar to the superheterodyne.

5. IMAGE INTERFERENCE.

For every setting of the local oscillator there are two signal frequencies which could produce the required i-f, one above and the other below the oscillator frequency. For example, if the i-f is 455 Kc/s and the oscillator frequency is 6455 Kc/s, signals at 6000 Kc/s and 6910 Kc/s both provide the correct i-f. One of these two signals is the required signal and the other is the interfering signal. The latter is called the *image frequency* and gives image or *second-channel interference*. This is one of the most common types of signal interference in superhets.

The ratio of the output of the receiver at the desired frequency to the output at the image frequency is called the *image ratio*. This ratio may be made large by using a preselector and an i-f such that no two stations differ in frequency by twice the i-f. Also, the higher the i-f the farther the image frequency is from the desired frequency and the more easily it is eliminated by the input tuned circuits. In high-frequency receivers it is sometimes necessary to use a much higher i-f in order to eliminate image interference. If, for example, a receiver uses an i-f of 200 Kc/s, stations which are 400 Kc/s apart will produce image interference. If, however, the i-f is raised to 500 Kc/s, stations must be 1000 Kc/s apart to produce image interference. Stations this far apart are readily separated by the input circuits of the receiver.

6. Heterodyning Between Stations Operating at Frequencies Differing by the I-F.

Two strong local stations, differing in frequency by the i-f, are sometimes observed to pass through the receiver even when the local oscillator is inoperative. In this case one station heterodynes with the other to produce the i-f. The trouble is reduced by improving the selectivity of the input, installing wave traps at the interfering frequencies or changing the i-f. The use of an i-f such as 456 or 465 Kc/s which is never the difference between two signal frequencies on the broadcast band is suggested.

7. Transmitters Operating at the I-F.

Any strong station operating at the i-f is permanently tuned in by the i-f amplifiers and therefore continually interferes with other stations. To eliminate this interference, good selectivity is necessary before the first detector and good shielding in the i-f stages. The use of wave traps and a different i-f are also possible methods of reducing the interference.

8. Harmonics of the Local Oscillator.

Harmonics of the local oscillator may mix with signals reaching the first detector to produce an i-f signal which is unwanted. For example, a signal at 700 Kc/s is to be received by a set having an i-f of 450 Kc/s. The local oscillator is set to operate at 1150 Kc/s. The 2nd harmonic of the l.o. has a frequency of 2300 Kc/s. Stations which transmit at frequencies of 2750 Kc/s and at 1850 Kc/s will produce i-f signals at 450 Kc/s. These stations may therefore interfere with the reception of the desired signal unless adequate selectivity is provided before the first detector.

If a transmitter is keyed sharply a certain amount of energy is radiated in sideband frequencies far from the carrier frequency. This energy, while small, may cause serious interference in neighbouring receivers, even when they are not tuned to the same frequency as the interfering transmitter. This can be eliminated only at the transmitter.

B. Remedies for Interference Between Signals

Reference has already been made to the advantages of using well shielded circuits in the r-f and i-f stages of receivers to eliminate certain types of interference. High selectivity in the r-f stages is achieved largely by using high-Q resonant circuits for tuning. The heterodyne principle itself provides most of the selectivity in the superhet receiver. In addition, the use of a Class A r-f amplifier before the first detector is a distinct advantage in attaining good selectivity. Such a circuit is called a *pre-selector*. It provides a boost in signal voltage before the frequency converter which is useful in eliminating cross modulation and harmonic interference. In

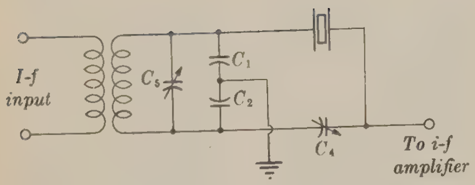


FIG. 9-48 A crystal-filter circuit.

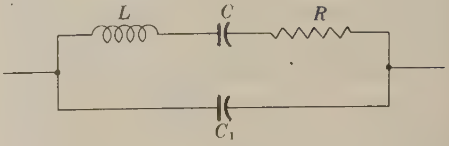


FIG. 9-49 Equivalent circuit of a crystal.

addition the pre-selector is useful in preventing radiation from the local oscillator of the receiver through the receiving aerial.

The proper choice of i-f plays a large part in reducing certain types of interference. As has been noted, a low i-f is advantageous in eliminating adjacent-channel interference whereas a high i-f is least likely to cause image interference. The actual choice of intermediate frequency is therefore a compromise between these two requirements. An i-f. of 456 Kc/s is generally used in broadcast receivers and is also widely adopted for communications equipment. A small shift above or below this frequency may usually be made to avoid interference in a particular receiver.

The use of crystal filters and wave traps to discriminate against interfering signals is discussed in the following sections.

9.17 Crystal filters. The performance of an i-f amplifier in providing a large output signal at the desired frequency and discriminating against signals at neighbouring frequencies is greatly improved by the use of a crystal filter circuit. An i-f crystal filter is shown in Fig. 9-48. The secondary winding of the i-f transformer is balanced to ground by the capacitors C_1 and C_2. One end of the secondary is connected to the grid of the following amplifier by the crystal. The other end of the i-f transformer is connected to the grid through C_4 which is called the *neutralizing condenser* or *phasing control*. The circuit is designed to provide two advantages; improved selectivity and single-signal reception. The operation of the circuit in attaining these advantages is described in the following paragraphs.

A. Selectivity

To understand the action of the circuit, it is first necessary to recall the characteristics of the crystal itself. The crystal is of the type used to control the frequency of an oscillator and has the electrical characteristics of a high-Q series resonant circuit. The equivalent circuit of the crystal is shown in Fig. 9-49. L, C and R represent the equivalent inductance, capacitance and resistance of the crystal itself while C_1 represents the capacitance of the crystal holder. The crystal is ground so that its resonant frequency is the same as the i-f of the receiver. At this frequency it operates as a small resistance and passes a high current to the grid circuit of the following amplifier. At frequencies off resonance, the impedance of the crystal circuit is high and the current falls to a low value. The phasing control is used to maintain the selectivity of the circuit at a high value. It is adjusted to balance out the capacitance provided by the crystal holder and thus to prevent bypassing undesired frequencies from the i-f transformer to the grid circuit of the following amplifier. Since the secondary winding of the i-f transformer is balanced to ground by C_1 and C_2, the voltages at its opposite ends are 180° out of phase with each other. One of these voltages is applied to the grid through the crystal holder capacitance and the other through the phasing control. When this control is adjusted so that C_4 equals the crystal holder capacitance, the out of phase voltages are equal and cancel each other. In this way C_4 prevents signals at undesired frequencies from reaching the grid circuit of the amplifier.

The selectivity of the coupling circuit may be varied by adjusting the selectivity control C_5. C_5 is normally adjusted to tune the secondary winding of the i-f transformer to resonance at the intermediate frequency. When this is done, the impedance of the secondary circuit is a maximum and is purely resistive. This resistance is in series with the crystal circuit and reduces its effective Q. If it is desired to increase the selectivity of the receiver, C_5 is detuned from its resonance setting. This reduces the impedance of the secondary circuit which is in series with the crystal and so the Q of the circuit is maintained at a high value. At the same time, the gain of the coupling circuit is decreased. That is, maximum gain and maximum selectivity cannot be obtained at the same setting of the selectivity control.

B. Single-Signal Reception

The phasing control also operates in conjunction with the crystal to provide adjustable rejection of interfering signals. This is readily understood by reference to a particular example. A receiver with i-f equal to 456 Kc/s is to discriminate between c-w signals differing in frequency by 1 Kc/s. That is, the receiver must discriminate between the desired signal and interfering signals which produce i-f's of 454, 455, 457, 458 Kc/s, etc. If the frequency of the b.f.o. is set at 457 Kc/s, a 1000 c/s audio note is produced in the output by the desired signal and by the signal producing an i-f of 458 Kc/s. The other signals are not objectionable because the one at 457 Kc/s produces zero beat while the others produce higher pitched audio notes that may be over-read. The interfering signal at 458 Kc/s is called the *audio image*. It may be eliminated by proper adjustment of the phasing control as follows.

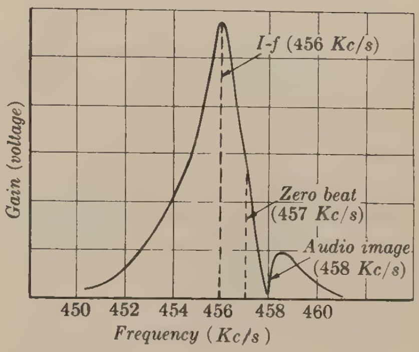


FIG. 9-50 Crystal-filter response for single-signal reception.

The phasing capacitor C_4 is varied slightly from the setting at which it balances the crystal holder capacitance. This leaves a small effective capacitance in parallel with the crystal circuit. The resonant frequency of this parallel circuit is slightly higher than the resonant frequency of the crystal itself. That is, the crystal must present an inductive reactance to resonate with the capacitive reactance of the crystal holder. To present an inductive reactance, the crystal must be operated above its resonant frequency.

By proper adjustment of the phasing capacitor, it is possible to set the resonant frequency of the parallel circuit at 458 Kc/s. At this frequency, the circuit presents a very high impedance and so the audio image is eliminated. The response curve of the crystal circuit under these conditions is shown in Fig. 9-50. In operation therefore, the b.f.o. is set to oscillate at a frequency above the i-f and the phasing control is adjusted to eliminate the audio image. This is called *single-signal reception.*

9.18 Wave traps. Interference caused by a strong signal at a fixed frequency

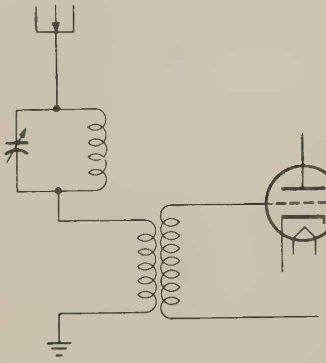


FIG. 9-51 Aerial circuit with wave trap.

may largely be eliminated by the use of *band-suppression filters* or *wave traps.* A simple wave trap which is effective in eliminating an undesired signal from a nearby transmitter is shown in Fig. 9-51. It consists of a high-Q parallel resonant circuit connected in the aerial circuit of the receiver. The circuit is tuned to resonance at the frequency of the interfering signal. It therefore presents a high impedance to currents at this frequency and a low impedance to other aerial currents. A large voltage is developed across the circuit at the resonant frequency and only a small voltage is provided across the primary of the aerial transformer. The interfering signal therefore becomes ineffective in later stages of the receiver. If the wave trap is sharply tuned, it has very little effect on the strength of the output signal at frequencies off resonance.

The wave trap may be connected in the cathode circuit of an r-f amplifier as shown in Fig. 9-57. The wave trap produces high-impedance degeneration at that frequency. Thus the gain of the stage is very small and the unwanted signal is eliminated.

If the i-f of the receiver is in the broadcast band it is usually necessary to include at least one wave trap in the receiver circuit to eliminate interference from stations in that band. Sometimes two or more wave traps are included in the same receiver. They may be tuned to the same frequency or to different interfering frequencies. The tuning may be effected by a variable capacitor or by permeability tuning of an iron-core inductor.

9.19 Noise. The term noise is used here to refer to a wide range of electrical interference other than that caused by unwanted signals. It includes spurious voltages from the atmosphere, from electrical equipment other than transmitters and from components within the receiver itself which mask the audio signal as it comes from the speaker.

A measure of the interference between signals and noise is given by the *signal-to-noise ratio.* Only if this ratio is sufficiently high is it possible to receive signals intelligibly. There are two obvious ways by which the ratio can be improved: by increasing the strength of the signals and by decreasing the strength of the noise. From a practical point of view the latter is the better method. With this object in view, the various types of interference encountered in radio receivers are considered with some of the remedies.

A. Atmospherics

This type of interference is caused by electrical disturbances in the atmosphere. The disturbances are sometimes violent as in the case of lightning or weak as in the case of a gradual discharge between cloud and earth. Static impulses, having components at all frequencies, cannot be tuned out and they affect reception over all bands. The impulse shocks the input circuit of the receiver into oscillation at its own resonant frequency and the receiver responds. These impulses are usually much stronger than signals from weak distant stations but last only a few thousandths of a second.

Static is worse during warm weather

than cool weather and worse at sunset than at other times of the day. Static that is characteristic of warm or changing weather produces almost continual crashing sounds. When an electrical storm is gathering a continuous hissing sound is produced in the receiver output, caused by steady electrical discharge through the aerial system. This effect is noticeable during snow storms as the small charges on the snow flakes are discharged through the aerial. Lightning discharges, either near at hand or far distant, cause loud crashes and intermittent rattling noises.

A high resistance, 100,000 ohms or more, connected across the input of the receiver by-passes more of the static than of the signal because the receiver is tuned to the signal. The use of such a resistor increases the signal-to-noise ratio under unfavourable conditions of reception.

B. Electrical Interference
(Man-Made Static)

Interference of this type may be caused by surges, sparks or any other sudden changes in current or voltage. The interference may reach the receiver in many ways:

(i) by direct radiation from the source to aerial
(ii) by conduction along the power lines to receiver
(iii) by radiation from the improperly shielded power lines to the aerial
(iv) by conduction along the ground.

Such disturbances may be caused by sparking at commutators in motors and generators, by switches, relays, keys, vibrators, thermostats, etc., which make and break the circuit. Leaky lightning arrestors, cracked insulators, loose wiring or even generator defects may also give rise to receiver noise. The best way to eliminate noise of this description is to suppress it at the source. Proper design of equipment and the use of filters should reduce the noise to the point where it can not obscure even the weakest signal.

C. Receiver Noise

In tracing the source of noise it should first be determined whether or not the noise originates outside of the receiver. Much of the noise present is traceable directly to the receiver operation. The following are some of the sources of interference in receivers.

1. THERMAL AGITATION (RESISTOR NOISE).

In a conductor the free electrons are moving in random directions. This motion causes small irregular e.m.f.'s to be developed across the terminals of the conductor. These e.m.f's increase as the resistance and temperature are raised. This effect is called *thermal agitation*. If the conductor is connected across the input to a receiver a hissing noise is heard in the receiver output. The e.m.f. appearing across the input tuned circuit of a receiver due to thermal agitation depends on the dynamic resistance of the parallel circuit, the temperature and the frequency band of reception. Thermal agitation and hence noise in the output, can be reduced by designing the input circuit to have a low dynamic resistance, keeping the temperature low and making the band-pass as narrow as the application permits.

2. VALVE NOISE.

Another component of the hiss noise in a receiver is developed in the valves. It is produced by the random emission of electrons from the cathode which causes an irregular component in the anode current. This is called the *shot effect*. Since valve noise is amplified throughout the receiver, the first valve contributes most of this type of noise. In gaseous valves, sudden discharges cause a large amount of noise to be generated.

3. MECHANICAL DEFECTS.

In volume controls a great deal of noise may be produced if the variation of resistance is made non-uniform by poor or dirty contacts. The voltage surges set up by abrupt changes in resistance cause the noise. On wire-wound controls the resistance always changes by steps so that noise is usually present when the control is varied. Similarly, poor solder joints may cause abrupt changes in resistance in a circuit under vibration with resultant voltage surges and noise.

4. MICROPHONICS.

A further source of noise arises from the fact that vibration of a valve may cause changes in the valve constants. This results in variation in the current through the valve which in turn produces noise. Noises of this type are called *microphonics*. To remedy this, a valve of rigid construction should be used and mounted in such a way as to minimize the effect of vibration.

D. Remedies

To obtain the best signal-to-noise ratio, the signal must be made as large as possible at the grid of the first valve. This involves the use of a high-Q input circuit and a valve with low inherent noise. The first stage of the receiver is the important one from the standpoint of signal-to-noise ratio. Noise generated in the later stages, while comparable to that generated in the first stage, is masked by the amplified noise from the first stage. At lower frequencies most of the receiver noise comes from thermal agitation in the aerial circuit. At higher frequencies, however, as the dimensions and impedance of the input circuit are reduced, valve noise becomes more predominant. Also the signal-to-noise ratio is usually low because high-Q circuits are difficult to design.

Low hissing background noise which arises from thermal agitation, valve noise and even atmospherics may be reduced to some extent by improving the selectivity of the receiver. Since noise of this type consists of random voltages over a wide range of frequencies, a reduction in the receiver band-pass serves to reduce the noise.

Noise which results from short sharp voltage impulses of large amplitude may be removed to a large extent by noise limiters. Several circuits used for this purpose are described in the following section.

9.20 Noise limiters and noise suppressors. Noise limiters are used to reduce large noise pulses which are often superimposed upon the signal. They do not remove the noise altogether, but merely keep it from reaching a value larger than that of the signal. The limiter usually follows the second detector and so operates on the a-f signal. In order to appreciate the operation of the limiter, it is necessary to examine the operation of the detector when the r-f signal contains noise pulses. Fig. 9-52 (a) and (b) show a diode detector circuit and typical waveforms of the r-f signal, current through the valve and a-f signal. The noise appears as sharp spikes of random occurrence and amplitude along the regular waveform. It should be noted that the rectifying action of the valve cuts off noise pulses which occur during the negative half cycles of the r-f signal. The noise pulses in the a-f output voltage are therefore restricted to those on the positive side of the signal. Consequently noise limiters are required to operate on noise pulses of one polarity only.

Two types of noise limiters are widely used, shunt and series limiters.

A. Shunt Limiter

The shunt limiter, Fig. 9-53 (a), consists of a biased diode V_2 in parallel with

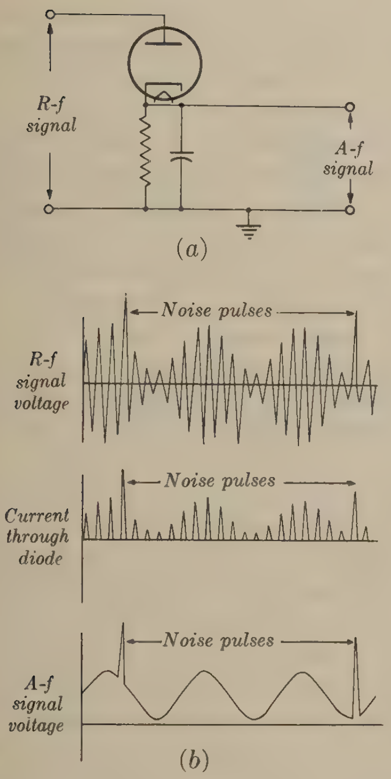


FIG. 9-52 Detection of r-f signal containing noise pulses (a) Diode detector circuit (b) Voltage and current waveforms.

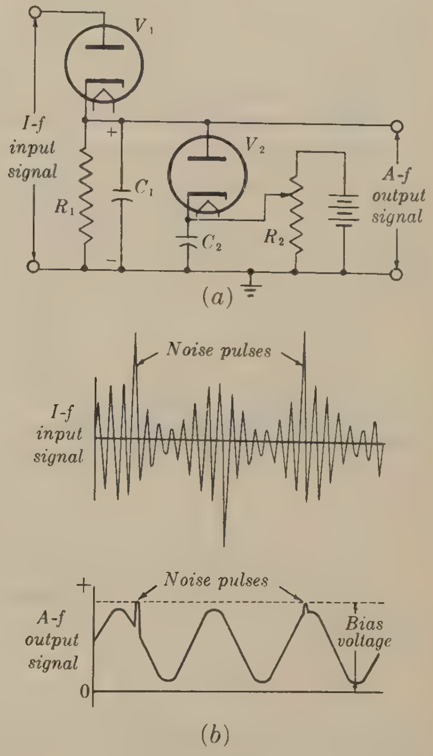


FIG. 9-53 Diode detector with shunt noise limiter (a) Circuit diagram (b) Effect of limiter on noise pulses.

the detector load R_1. The anode of the limiter is connected to the upper end of R_1 which is at a positive d-c potential due to the d-c component of rectified current flowing through V_1. The cathode of the limiter is connected to the potentiometer R_2 which receives a positive potential from the bias battery. The tap on R_2 is set to make the cathode of V_2 slightly more positive than the maximum positive potential attained at the cathode of V_1 during the a-f cycle. Hence, V_2 is normally biased off and presents a high impedance in parallel with the detector load resistor R_1. The limiter therefore has no effect on a-f signals of normal amplitude. However, when a noise pulse of positive polarity in the i-f signal is applied to V_1, the valve conducts more heavily causing a corresponding rise in the cathode potential of V_1. The anode of V_2 is therefore made positive to its cathode. The limiter conducts current and presents a low impedance in parallel with R_1 during this time. The voltage across V_2 is almost zero when conducting and C_2 bypasses R_2 during the pulse. In effect then, the noise pulse is short-circuited by V_2 and the waveform is clipped off at the bias level as shown in the diagram Fig. 9-53 (b).

B. Series Limiter

This type of detector, Fig. 9-54 (a), uses a diode as the coupling device from the detector to the first a-f amplifier. Its operation differs from that of the shunt limiter in that V_2 conducts during the reception of signals of normal ampli-

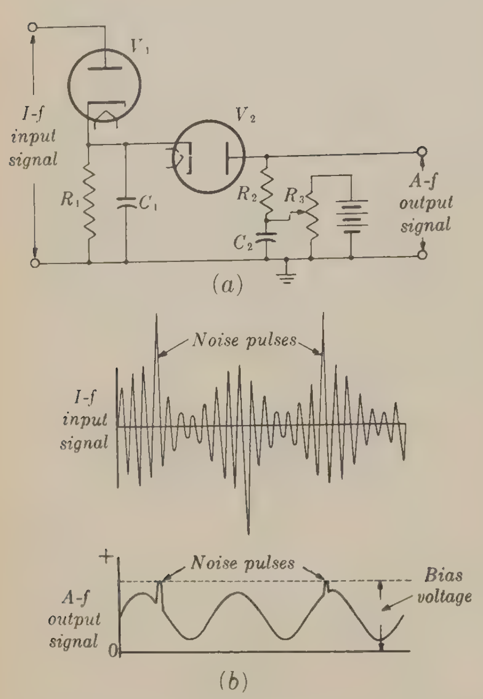


Fig.9-54 Diode detector with series noise limiter (a) Circuit diagram (b) Effect of limiter on noise pulses.

tude and is cut off during noise pulses. A more detailed description of the circuit follows. The cathode of V_2 is connected to the upper end of the load resistor R_1. The anode of V_2 is connected through R_2 to a positive point on the potentiometer R_3. When no signal exists across R_1, the limiter conducts a small current which maintains the anode of V_2 positive to ground. When the signal causes the cathode of V_2 to vary in potential, a varying current flows through the valve. This develops the varying a-f voltage across R_2 which is the output signal. When a noise pulse is received, the cathode potential of V_2 becomes more positive than the anode potential. The diode stops conducting for the duration of the pulse and the waveform is clipped off at the bias level, Fig. 9-54 (b).

In practice it is not usual to use a battery to produce the bias as indicated in the above diagrams. The battery may be replaced by a voltage divider or by a resistor-capacitor combination. Fig. 9-55 shows a limiter circuit in which the bias is obtained by the latter method. The voltage across the load R_1 appears across R_2 and C_2 in series. The output voltage is taken between the tap on R_1 and ground. It is of negative polarity and is applied to the anode of V_2. The d-c voltage developed across C_2 is applied to the cathode of V_2. The time constant R_2C_2 is sufficiently large to maintain only the d-c component of the detector output voltage across C_2. The limiter normally conducts current because its cathode is at a more negative potential than its anode. The a-f signal is therefore coupled through V_2 and C_3 to the following stage. Noise pulses of positive polarity in the i-f signal are not passed by the detector. When a noise pulse of negative polarity is received, the cathode of V_2 is made more negative than the anode and so the valve stops conducting. Consequently the amplitude of noise pulses of short duration is limited to the d-c voltage across C_2. If, however, the duration of the noise pulse is long enough to allow C_2 to charge through R_2, V_2 conducts again. The noise pulse is then coupled to the following circuit and the limiter becomes ineffective.

C. *Noise Suppressors*

The sensitivity of a receiver provided

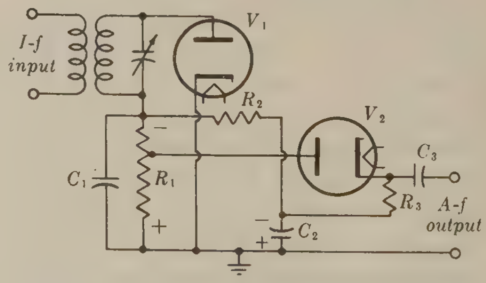


Fig.9-55 A diode detector and limiter circuit.

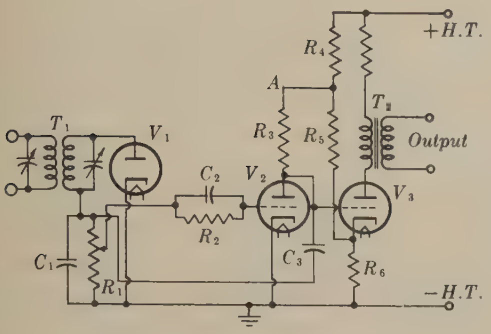


FIG.9-56 A noise-suppressor circuit.

with a.v.c. varies with the strength of the received signals. When the set is being tuned in search of a station, the sensitivity is a maximum. As soon as a station is tuned in, the a.v.c. bias increases and the sensitivity is reduced. During the searching process, when the sensitivity is a maximum, the noise level may become so large as to be disagreeable. A *noise suppressor* or *squelch circuit* may be used to eliminate the noise between stations.

A noise suppressor circuit is shown in Fig. 9-56. V_1, V_2 and V_3 are respectively the diode detector, noise suppressor and a-f voltage amplifier valves. The suppressor circuit permits the amplifier to operate when a signal is being received and turns it off when no signal is tuned in. A detailed description of the circuit follows.

The a-f output of the detector is developed across the load resistor R_1. This signal is coupled through C_3 to the grid of the voltage amplifier. The d-c component of rectified current through the detector develops a d-c voltage of negative polarity at the upper end of R_1. A part of this voltage is tapped off through the variable tap of R_1 to the grid of V_2. This voltage is filtered by R_2C_2, a low-pass filter to remove the a-f variations in the signal. Hence a negative d-c voltage which is proportional to the amplitude of the received signal is applied to the grid of V_2. This voltage is ordinarily large enough to bias V_2 beyond cut-off when a station is tuned in. In effect, therefore, V_2 is removed from the detector-amplifier circuit. The cathode potential of V_3 is positive to ground as it is connected to the voltage divider R_4, R_5, R_6 across the H.T. supply. Automatic bias is developed by C_3R_3 in the grid circuit and the valve acts as an amplifier producing an output signal in T_2.

When no signal is applied to the detector from T_1 the d-c voltage across R_1 disappears. The grid of V_2 is therefore at cathode potential and so the valve conducts heavily. The current flowing through the valve develops a large p.d. across R_3 so that the anode potential of V_2 and the grid potential of V_3 are at a low positive value. The cathode potential of V_3 is still positive from the voltage divider R_4, R_5, R_6. Under these conditions the grid is sufficiently negative to the cathode to cut off the current through V_3 and render the amplifier inoperative. Noise signals from the detector are therefore prevented from reaching the final output of the set and are not heard in the speaker.

9.21 A superheterodyne receiver circuit. Illustrated in Fig. 9-57 is the circuit of a Service superheterodyne receiver capable of c-w, m-c-w and r-t reception. This receiver is provided with several features not ordinarily found in non-Service receivers. Two stages of r-f pre-selection are used to ensure satisfactory image ratios; high-gain aerial coils provide a maximum signal-to-noise ratio. A crystal filter in the i-f stages permits the selectivity to be varied over a wide range. If desired, spot-frequency reception may be used. This is made possible by the use of plug-in crystals which control the frequency of the local oscillator. Adequate volume control is provided by a sensitivity control as well as a

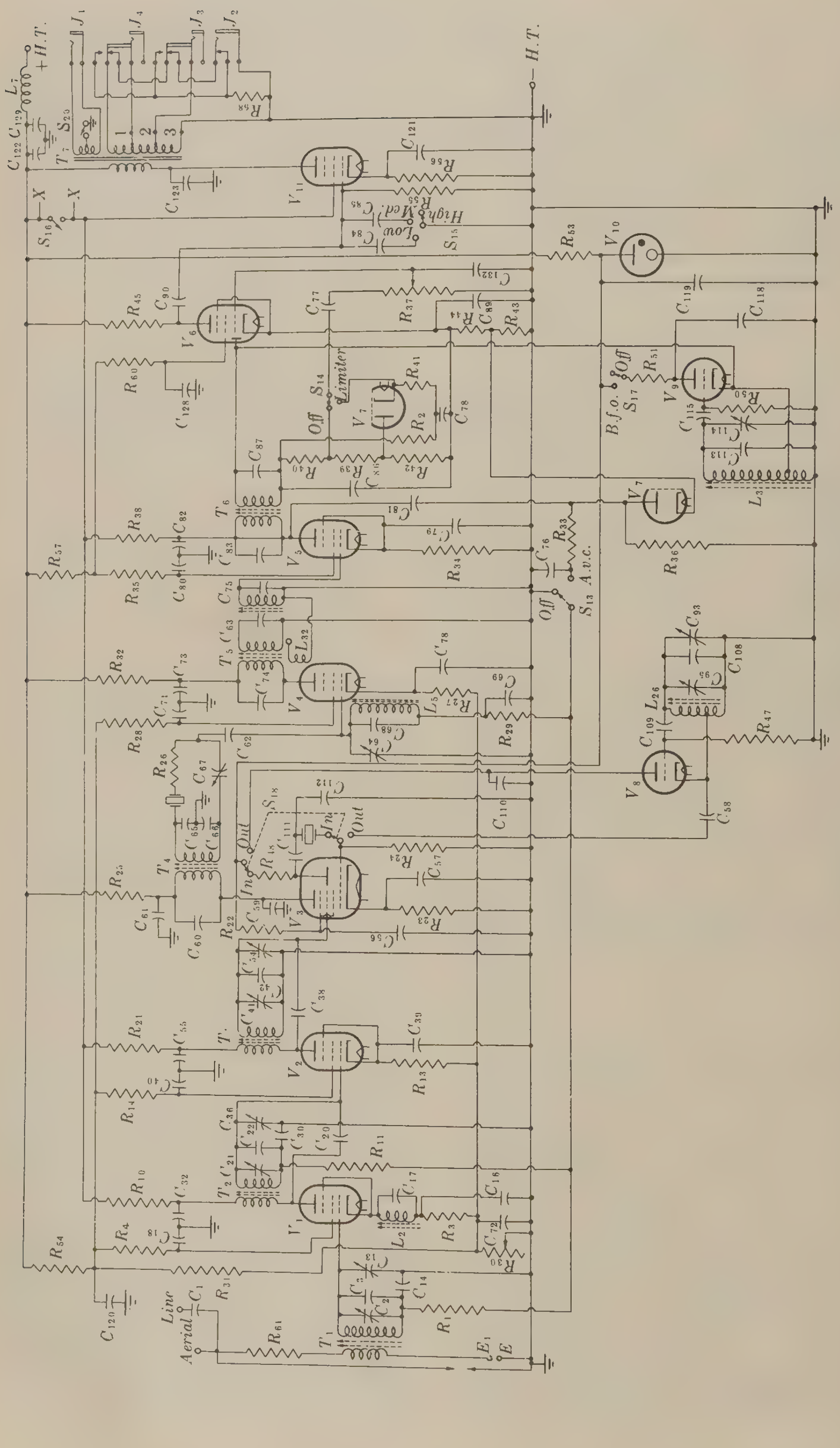


Fig. 9-57 A superheterodyne receiver circuit.

gain control. A noise limiter is present to provide satisfactory reception under adverse noise conditions. Temperature-compensating components make the receiver almost independent of temperature variations.

The tuned r-f circuits for only one frequency band are shown in Fig. 9-57. In practice, five other sets of tuned circuits are available to provide reception from low to high radio frequencies.

A. *Aerial Circuit*

A single-wire aerial may be used with the receiver. If a transmission line is employed, connection is made through C_1 which couples the input to the aerial input transformer. This capacitor is necessary in order that the tracking of the first stage is not mis-aligned when switching from single-wire aerial to transmission line input. R_{61} is a non-inductive resistor to reduce re-radiation to a safe level. To prevent damage to the r-f coils from powerful signals, a gas-filled discharge gap is situated in the aerial circuit. This gap breaks down at voltages exceeding 100 volts and shorts the aerial to ground to protect the r-f coil. Signals are coupled from the aerial circuit through the r-f transformer T_1 to the grid circuit of the first r-f amplifier. The terminals E and E_1 are connected together for a straight-wire aerial or a coaxial transmission line input. If the input is from a balanced line, one side of the line is connected to E_1, and E and E_1 are not connected.

B. *First R-F Amplifier*

This circuit is a tuned r-f voltage amplifier, operated Class A. A variable-mu pentode V_1 is used to allow a.v.c. to operate. The tuned input circuit consists of the secondary of the r-f transformer T_1 in conjunction with C_{13}, which is one section of the main tuning capacitor and is ganged with C_{36}, C_{54} and C_{93}. Across C_{13} is connected a negative-temperature-coefficient capacitor C_3 to compensate for changes in capacitance in the other capacitors with changing temperature. C_2 is a trimmer capacitor for tracking of the dial at the high-frequency end. R_1 and C_{14} are the a.v.c. decoupling resistor and capacitor. In the cathode lead, L_2 and C_{17} form a wave trap which is permeability tuned to the i-f. The wave trap prevents any signal near the i-f from passing through the receiver. R_3, C_{16} form a cathode bias resistor and by-pass capacitor. The R-F GAIN control is R_{30} which varies the bias on the cathodes of the first and second r-f amplifiers and the first i-f amplifier. R_{30}, a rheostat, is part of a voltage divider across the H.T. made up of R_{54}, R_{31} and R_{30}. C_{120} is an H.T. filter capacitor, C_{72} is an r-f by-pass capacitor. The anode load is T_2, an r-f coupling transformer with a tuned secondary circuit. C_{20} across this transformer provides additional coupling to make up for losses at high frequencies. R_{10} and C_{32} are the anode decoupling resistor and capacitor. R_4 and C_{18} are the screen voltage dropping resistor and screen by-pass capacitor.

C. *Second R-F Amplifier*

The second r-f amplifier V_2 is almost exactly the same as the first r-f amplifier. It is a Class A tuned r-f voltage amplifier whose output is coupled to the frequency converter through the transformer T_3 and by C_{38}. All other components have the same functions as the corresponding components in the first r-f amplifier stage.

D. *Frequency Converter*

1. MIXER AND FIRST DETECTOR

The mixer and first detector valve is the hexode section of V_3. The signal is applied to the third grid of the hexode and is detected by the anode-bend operation of this section. The local oscillator voltage is coupled to the first grid. Since both of these grids control the electron stream, heterodyning occurs. The anode load, consisting of C_{60} and the primary of T_4, is tuned to the i-f and the r-f signal is decoupled by R_{25} and C_{61}. C_{59} is inserted to help provide the desired selectivity. R_{22} and C_{56} are the voltage dropping resistor and r-f by-pass capacitor for the second and fourth grids. R_{23} and C_{57} are the cathode bias resistor and by-pass capacitor.

2. Crystal-Controlled Local Oscillator.

This local oscillator is a Pierce crystal oscillator that is used for spot-frequency reception only. With the Crystal Oscillator switch S_{18} in the In position, pole 1 connects the crystal feedback to the grid of the triode section of V_3 and pole 2 connects the H.T. to the anode of the triode section to operate the oscillator. With S_{18} in the Out position, pole 1 connects the tuned local oscillator output to the mixer and pole 2 connects the H.T. to the anode of V_8, to set the tuned local oscillator in operation. To receive a signal using this crystal oscillator, a crystal, whose frequency is higher or lower than the signal frequency by an amount equal to the i-f is plugged into the converter crystal socket and S_{18} switched to In. C_{111} is an r-f coupling and d-c blocking capacitor which together with C_{112} provides the correct feedback to maintain oscillations. Automatic bias for the oscillator is provided by R_{24} and the crystal holder capacitance. R_{48} is the anode load resistor.

3. Tuned Local Oscillator.

For continuous reception over all bands, the crystal-controlled l.o. is switched off. This allows the tuned l.o., using V_8 a modified Hartley oscillator, to operate. The frequency of this oscillator is above the signal frequency by an amount equal to the i-f. The output from this circuit is coupled through C_{58} to the mixer valve. The tuned circuit consists of L_{26} and C_{93}, C_{95} and C_{108}. L_{26} is a permeability-tuned inductor for tracking at the low-frequency end of the dial. C_{93} is part of the main ganged capacitor. C_{95} is a trimmer for tracking at the high-frequency end of the dial. C_{108} is a negative-temperature-coefficient capacitor. C_{109} and R_{47} provide automatic grid bias. C_{110} is the r-f by-pass capacitor for the anode.

E. First I-F Amplifier

The output from the mixer, V_3, is coupled by T_4 through the selectivity control circuit (discussed in sub-section *F*) and C_{62} to the grid of the first tuned i-f amplifier which is operated Class A. L_5 and C_{68} form the tuned input circuit providing a high impedance from grid to ground. It is tuned to the i-f. R_{29} and C_{69} are the a.v.c. decoupling resistor and capacitor. R_{27} and C_{78} are the cathode bias resistor and by-pass capacitor. The anode load is formed by C_{74} and the primary of T_5. R_{32} and C_{73} are the anode decoupling resistor and capacitor. R_{28} and C_{71} are the screen voltage dropping resistor and by-pass capacitor. T_5 couples the output of the first i-f amplifier to the grid of the second i-f amplifier.

F. Selectivity Control

The complete Selectivity control switch has four positions which provide for four degrees of selectivity.

1. Position 1.

In this position, illustrated in Fig. 9-58, the mixer is coupled to the first i-f amplifier by capacitor C_{59} and degeneration allowed in the coupling to the second i-f amplifier V_5. The degeneration is accomplished by using the auxiliary coil L_{32}. This provides a wide flat band-pass of approximately 8 Kc/s and is used for high-fidelity r-t reception when interference is not present. L_5 and

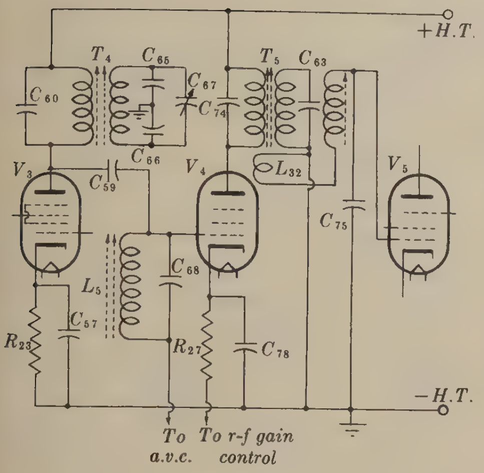


FIG.9-58 Circuit arrangement for position 1 of the selectivity control.

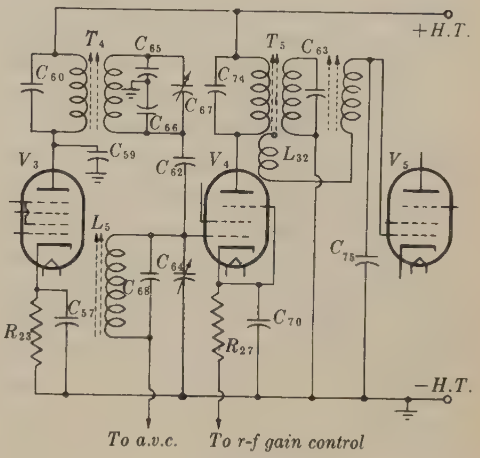


FIG.9-59 Circuit arrangement for position 2 of the selectivity control.

C_{68}, the tuned grid circuit of V_4, are detuned to obtain a wide flat-topped band-pass. The secondary of T_4, tuned by C_{65} and C_{66} in series with each other and the small capacitor C_{67}, helps to broaden the frequency response.

2. POSITION 2.

The mixer V_3 is coupled to the first i-f amplifier V_4 by tuned-transformer coupling T_4 (Fig 9-59). This provides a narrower band-pass of about 4 Kc/s. C_{64} is a fine tuning capacitor in parallel with C_{68} so that the tuned grid circuit of V_4 is brought back into tune on bands 2, 3 and 4. L_{32} is not used in position 2.

3. POSITION 3.

The same coupling from mixer to first i-f amplifier is used as in position 2. In addition, a capacitive bridge network containing a crystal is connected from the secondary of T_4 to the grid of V_4 (Fig. 9-57). This bridge consists of C_{65}, C_{66}, C_{67} and the crystal-holder-plate capacitance. The centre point between C_{65} and C_{66} is grounded to provide out-of-phase voltages for neutralizing the capacitance of the crystal holder plates. C_{67}, a phasing capacitor, is adjusted so that the voltage across it is equal to the voltage across the crystal holder plates. Since these voltages are made equal and in opposite phase, they cancel out leaving only the crystal and resistor for the coupling path from the mixer to the first i-f amplifier. With this arrangement, the band-pass is approximately 500 c/s and is used only when the interference is very strong.

4. POSITION 4.

In Position 4, the resistor R_{26} is shorted out, the other connections being essentially the same as in position 3. With R_{26} no longer in series with the crystal, the maximum selectivity of the crystal is employed to narrow the band-pass to about 100 c/s. This position is used only on c-w reception under the most extreme interference conditions. In position 4, L_{32} is again in use as in position 1.

The presence of a second tuned secondary circuit in T_5 using C_{63} increases the attenuation of all frequencies to which the circuit is not tuned. As a result, in all four positions of the selectivity control, the response curve has steeper sides.

G. Second I-F Amplifier

The second i-f amplifier is a Class A i-f voltage amplifier with a tuned input consisting of the secondary of T_5 and

C_{75} and a tuned output consisting of the primary of T_6 and C_{83}. V_5 is a variable-mu pentode. However the gain of this stage is not controlled by either the r-f gain control or the a.v.c. voltage. The output of this stage is coupled through T_6 to the second detector and through C_{81} to the delayed a.v.c. circuit. R_{34} and C_{79} form the cathode bias resistor and by-pass capacitor. R_{38} and C_{82} provide anode decoupling. R_{57} is a voltage-dropping resistor for the screen of V_5 and V_6. R_{35} and C_{80} provide screen decoupling.

H. Delayed A.V.C.

The output from V_5 is capacitively coupled through C_{81} to the anode of one section of a double diode V_7. The negative d-c voltage for a.v.c. does not appear across R_{36} unless the signal drives the anode of this diode more positive than the positive bias on the cathode from R_{43}. When the diode does conduct current flows through R_{36} to produce a negative d-c voltage at its upper end which is filtered by R_{33} and C_{76}. When S_{13}, the A.V.C.-OFF switch, is in the A.V.C. position, the a.v.c. voltage is applied to the grids of V_1, V_2 and V_4.

I. Second Detector

The output from V_5 is applied to the diode section of the diode pentode V_6. The input tuned circuit consists of the secondary of T_6 and C_{87}. The load of the diode detector is R_{40}, R_{39} and R_{42} in series. Filtering is provided by C_{86}. The output taken across R_{39} and R_{42} is coupled through C_{77} to the a-f voltage amplifiers when the noise LIMITER switch S_{14} is in the OFF position. When the noise limiter is turned on, the output passes through the limiter valve V_7.

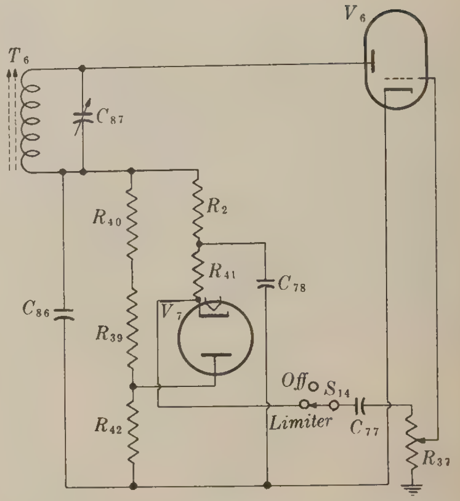


FIG. 9-60 Diode detector and noise limiter.

J. Noise Limiter

The noise limiter circuit is redrawn in Fig. 9-60. The noise limiter valve is one section of the double diode V_7. This section is connected in series with R_2 and R_{41}; this series combination is connected in parallel with R_{39} and R_{40}. Under normal operation the diode conducts current since its anode is at a more positive potential than its cathode. C_{78}, which is connected across R_{41}, the diode and R_{42}, is charged to the voltage across these components. To change its voltage C_{78} must charge through R_2; the time constant of C_{78} and R_2 is one-tenth of a second. The a-f voltage across R_{42} causes the anode voltage and hence the anode current of V_7 to vary at the audio rate. This varying current through R_{41} causes an a-f voltage across R_{41} which is coupled through C_{77} to the grid of V_6 when S_{14} is in the LIMITER position.

If a large noise pulse enters the detector it appears across the load R_{40}, R_{39} and R_{42}. For a short interval the cathode of V_7 is held constant in potential while C_{78} charges through R_2. For the duration of the pulse, however, the anode is driven more negative than the cathode and the diode is cut off. Thus the output is limited to the voltage across R_{41} and C_{78}. However, if the noise is not of short duration, C_{78} charges up, the diode conducts and the noise output is applied to the a-f stages. Thus the noise limiter is effective only for noise pulses of high amplitude and short duration.

K. Beat-Frequency Oscillator

For the reception of c-w signals, the B.F.O.-OFF switch S_{17} is turned to the B.F.O. position. The b.f.o. is a modified Hartley oscillator whose tuned circuit is L_3 and C_{113} and whose output is taken from the cathode lead of V_9 and applied to the second detector anode. The output from the b.f.o. heterodynes with the signal to produce a beat-frequency signal in the audio range. C_{113} is the basic capacitor in the tuned circuit while C_{114} is the variable tuning capacitor which varies the frequency of the b.f.o. slightly to vary the pitch of the audio note produced in the phones or speaker. C_{115} and R_{50} provide automatic grid-leak bias. R_{51} and C_{118} provide anode decoupling.

L. A-F Voltage Amplifier

The output of the detector or noise limiter is coupled through C_{77} and applied across the potentiometer R_{37}. This resistor serves as the A-F GAIN CONTROL, its output being coupled to the grid of the pentode section of V_6. This section is the a-f voltage amplifier valve which is operated Class A by the cathode bias across R_{44} and R_{43}, by-passed by C_{89}. The a-f voltage amplifier is resistance-capacitance coupled to the a-f power amplifier. R_{45} is the anode load resistor and C_{90} the coupling capacitor. C_{132} is an r-f by-pass capacitor. R_{60} and C_{128} provide screen decoupling.

M. A-F Power Amplifier

The input to the a-f power amplifier is coupled to the grid of V_{11}. The TONE control is situated in this grid circuit. The two capacitors C_{84} and C_{85} with the three-position switch S_{15} afford three different degrees of bass boosting when required. The capacitors by-pass the grid-leak resistor R_{55} for high audio frequencies, C_{84} being the larger capacitor. The a-f power amplifier is operated Class A and transformer coupled by T_7 to headphones or loudspeaker. The primary of T_7 is by-passed by C_{123} to improve the high-frequency response of the transformer. The secondary of T_7 has two windings. One of these matches the output to a 500-ohm line via the jack J_1. S_{20} is a shorting bar used to ground the centre tap of this secondary when a balanced line is used. The other secondary is tapped so that it may be used for matching to a loudspeaker via J_2, low-resistance phones via J_3 or high-resistance phones via J_4. R_{58} is an output matching resistor which is connected in place of the speaker when the phones are in use. For the loudspeaker, the whole secondary winding is used; for the high-resistance phones, part 1 to 3 is used; for the low-resistance phones, part 1 to 2 is used. R_{56} and C_{121} provide cathode bias. L_7, C_{122}, C_{129} provide H.T. decoupling. S_{16} is the SEND-RECEIVE switch which is closed when the receiver is in operation.

N. Voltage Regulator

V_{10} is a voltage regulator in series with the voltage dropping resistor R_{53} across the H.T. supply. V_{10} supplies a very steady voltage for the screen of the mixer V_3, the anode of the crystal local oscillator V_3 and the anodes of the local oscillator V_8 and b.f.o. V_9 for stability. C_{119} is an a-c filter across V_{10}.

O. Power Supply

The power supply (Fig. 9-61) is a full-wave rectifier circuit using a double diode V_{12}. The 115V a-c input to the transformer T_8 is applied via the ON-OFF switch S_{19}. L_8, L_9, C_{137}, C_{138} provide a two-section choke-input filter. R_{63}

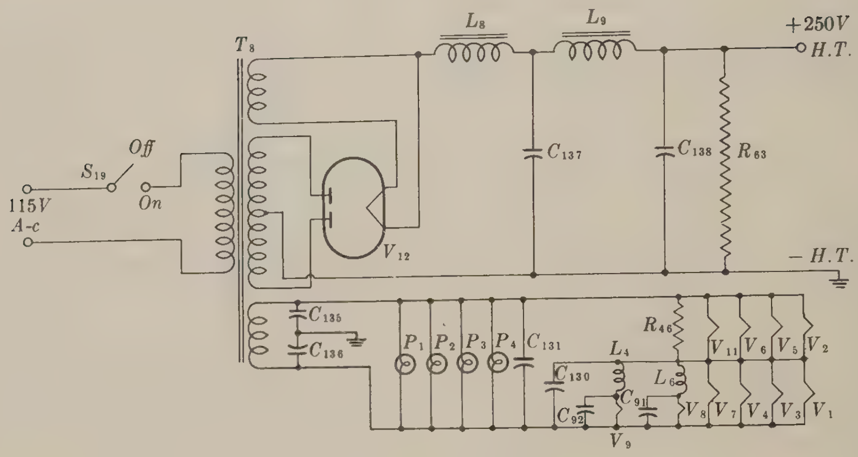


FIG. 9-61 Receiver power supply.

is a bleeder resistor. The d-c output is 250V. The filament secondary winding on T_8 provides 12V a-c for the pilot lamps and filaments of the receiver. C_{135}, C_{136} form a hash filter. C_{131} is an r-f by-pass across the 12V secondary. L_4 and C_{92} decouple the filament of the b.f.o. valve V_9. L_6 and C_{91} serve the same purpose for the local oscillator valve V_8 filament. C_{130} is a filament by-pass. R_{46} is a voltage dropping resistor.

9.22 Questions and problems.

1. Make a block diagram of a simple receiver and state briefly the purpose of each block.
2. Make graphical diagrams to show the waveform of the input voltage of each stage.
3. Name five requirements of a good communications receiver and explain the meaning of each briefly.
4. What circuits contribute most to the selectivity of the receiver?
5. Upon what factors does the sensitivity of the set depend?
6. State two causes of instability in a receiver.
7. Name three types of receiver signals and sketch the waveform of each.
8. What is the chief advantage of transmission by c-w signals?
9. What is the chief disadvantage to transmission by r-t signals?
10. What is the heterodyne principle?
11. What is meant by the beat frequency? What beat frequency is produced when a 2460 Kc/s wave heterodynes with one of 2348 Kc/s?
12. When receiving a c-w signal, how can the frequency of the a-f output be changed?
13. What is meant by the term zero beat?
14. When receiving a signal of 403.5 Kc/s, there are two interfering signals, one at 403.0 Kc/s and the other at 405 Kc/s. Explain how the b.f.o. may be adjusted to eliminate one interfering signal and render the other ineffective.
15. State two advantages of regenerative detection over non-regenerative detection.

16. Draw a regenerative detector circuit. Explain the action of the control used to vary the amount of regeneration.
17. Why must the regeneration be variable?
18. How may a regenerative detector be used to receive a c-w signal? What name is given to the detector when used in this manner?
19. What advantage does the superregenerative receiver have over the regenerative type?
20. What is the purpose of a quench oscillator?
21. Make a circuit diagram of a self-quenching receiver and state the components which provide the quenching action.
22. What is the purpose of a squegger? How does it differ in operation from the self-quenching receiver?
23. What is the superheterodyne principle? State two advantages of a superhet over a t-r-f receiver.
24. Make a block diagram of a superhet receiver. Sketch the waveform of the input voltage of each stage.
25. What is the function of the local oscillator? What kind of output does it produce?
26. Explain how the intermediate frequency is maintained constant for all r-f signals received.
27. Why must the i-f amplifiers in a superhet receiver amplify over a narrow band of frequencies rather than at just one frequency?
28. What are the two fundamental processes in frequency conversion?
29. Name three types of frequency converter circuit.
30. What is a pentagrid valve? Draw its circuit symbol.
31. State the purpose of each grid in the pentagrid valve of Fig. 9-21.
32. Draw the circuit symbol of a triode-hexode valve. State the purpose of each grid when the valve is used in a frequency converter circuit.
33. State two methods of controlling the gain of an amplifier with constant input voltage.
34. Explain how changing the bias may be used to control gain. What type of valve is used for this purpose?
35. Make a diagram of a volume control which is used to vary the amplitude of the input signal.
36. What is the meaning of the term logarithmic taper applied to a volume control? Why is taper used?
37. Explain how automatic volume control is used to prevent fading.
38. Make a diagram of a simple a.v.c. circuit and explain how the control voltage is developed.
39. Why is it necessary to filter the detector output before applying it as a.v.c. voltage?
40. What advantage does delayed a.v.c. have over simple a.v.c.? To what does the delay refer?
41. Draw a circuit diagram of a detector that provides delayed a.v.c.
42. What voltage is amplified by an amplified a.v.c. circuit?
43. Illustrate by diagrams the difference between series and parallel feed of a.v.c.
44. Account for the fact that an amplifier with a.v.c. amplifies weak signals to a greater degree than an amplifier with no a.v.c.
45. Describe the construction of a tuning-indicator valve.
46. Describe and explain the appearance of the target of a tuning-indicator valve
    (a) when no voltage is applied to the grid
    (b) when a negative a.v.c. voltage is applied to the grid.

47. Make a diagram of an S meter circuit with forward-reading meter. Explain its operation when the strength of the received signal increases.

48. In what part of a circuit is a tone control usually connected? Illustrate by a diagram.

49. Explain the operation of the tone control
    (a) when the resistance is a maximum
    (b) when the resistance is a minimum.

50. What is meant by bass boosting? Show by a graph the effect of bass boosting on the frequency response of the output of an amplifier.

51. List six types of interference between signals. Which of these is reduced by increasing selectivity? by raising the i-f? by lowering the i-f?

52. Explain how a crystal filter may be used to improve the selectivity of a receiver. Why is it used in the i-f rather than the r-f stages?

53. How is the phasing control used to improve the selectivity of the crystal filter circuit?

54. What is meant by the terms audio image and single-signal reception?

55. Describe the operation of the phasing control in eliminating the audio image.

56. What is the purpose of a wave trap? Why does it not reject all signals?

57. State three types of noise in receivers.

58. Explain the meanings of the terms atmospherics, microphonics, thermal agitation, shot effect.

59. How does a reduction in band-pass of a receiver improve the signal-to-noise ratio?

60. Draw a detector circuit which includes a series noise limiter and explain briefly the action of the circuit when a large noise pulse enters the receiver.

61. What is the function of a noise suppressor? Draw a circuit diagram and explain the operation of a noise suppressor.

62. In the circuit of Fig. 9-57, what is the function of each of the following components: T_1, C_2, C_3, R_{61}, R_{30}, R_3, L_2 and C_{17}, R_{10} and C_{32}, C_{13}, T_3, C_{58}, C_{65} and C_{66}, C_{67}, C_{111} and C_{112}, R_{26}, S_{13}, S_{18}, C_{86}, C_{78}, R_{37}, R_{53}, C_{85}, R_{58}, S_{17}, C_{123}?

63. Explain the operation of the spark gap in the aerial circuit of Fig. 9-57.

64. Draw the mixer and local oscillator circuit of Fig. 9-57 when S_{18} is in
    (a) the IN position
    (b) the OUT position.

65. Explain the operation of the selectivity control for each of its four positions (Fig. 9-57).

66. Explain the operation of the diode limiter V_7 in Fig. 9-57.

67. Explain how delayed a.v.c. is obtained in the circuit of Fig. 9-57.

68. Explain the operation of the b.f.o. circuit V_9 in Fig. 9-57 and give the purpose of the circuit.

69. Explain the use of the tone control switch S_{15} in each of its three positions in Fig. 9-57.

70. Give the purpose of S_{20} in Fig. 9-57.

71. Redraw the secondary circuit of T_7 for
    (a) loudspeaker
    (b) low-resistance phones
    (c) high-resistance phones.

72. Explain the purpose of the following components in Fig. 9-61: C_{135} and C_{136}, L_4 and C_{92}, R_{46}.

CHAPTER X

TRANSMISSION LINES

The basic radio application of a transmission line is to efficiently transfer energy from a transmitter to an aerial so that maximum radiation takes place from the aerial. As a result, before beginning the study of transmission lines it is of first importance to understand the function of an aerial and the manner in which an aerial radiates its energy into free space.

10.1 Function of the aerial. The function of an aerial is to efficiently radiate energy from a transmitter into space in the form of electromagnetic waves; or to efficiently intercept energy for a receiver from electromagnetic waves in space. The specific requirements for an efficient transmitting aerial at any frequency are the same for an efficient receiving aerial at the same frequency, so that any discussion of transmitting aerials will apply equally to receiving aerials.

10.2 Radiation. Consider a simple vertical transmitting aerial as in Fig. 10-1 (a). If coupled to a transmitter, an alternating radio-frequency current flows in this aerial and since the wire has a certain capacitance to earth, it forms a capacitor with the earth and assumes alternately a positive and negative charge. Thus at any instant there will be lines of electric force between the aerial and earth. These lines of force are bowed out due to their mutual repulsion. When the current in the aerial reaches a maximum the *electric field* is at a minimum. As the current in the circuit decreases the electric field increases, becoming a maximum when the current is zero. An instant later when the current reverses, the *lines of electric force* build up again with opposite polarity. Fig. 10-1 (a) represents this effect with the arrows indicating direction of the electric field. Thus the electric field is in the vertical plane.

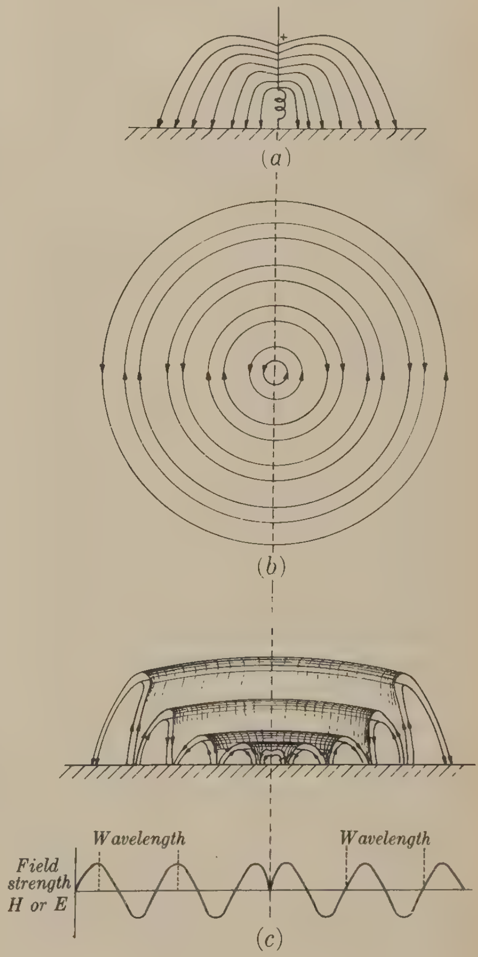


FIG.10-1 Radiation from a simple vertical aerial (a) Static electric field (b) Travelling magnetic field (c) Travelling electric field (d) Variation of field strength.

Besides this electric field there is a *magnetic field* due to the current in the aerial. Hence as the current increases the magnetic field increases and as the current decreases the magnetic field decreases. This field is represented in the conventional manner by lines of force forming concentric circles about the aerial, Fig. 10-1 (b). The figure represents a view looking down upon the aerial and the magnetic lines of force are in the horizontal plane represented by the plane of the paper. Only the lines of force in one plane near the earth are shown.

When the current in the aerial reaches a maximum the magnetic field is at a maximum. As the current in the circuit decreases the field decreases returning energy to the circuit. However, the magnetic field does not collapse in time with the current. That is, when the current starts to decrease the field immediately next to the wire starts to decrease but the more distant parts of the field do not begin to decrease until a short interval later. It takes a definite time for the effect of the decreasing current to make itself felt in the space outside it, and the farther a point is from the aerial the greater is the time lag between the decreasing current in the aerial and the decreasing field at the point. Thus the inside flux ring will start to collapse a small fraction of a second after the current has started to decrease. The second flux ring will start to collapse a small fraction of a second later than the first, and so on throughout the field until the outermost ring is reached. In this way the field does not collapse in time with the current but lags behind the current so that when the current reaches zero there is still a magnetic field outside the aerial.

The current in the wire does not remain at zero but reverses and begins to increase in the opposite direction and a new magnetic field builds up opposite in direction to the first field. This new field pushes outwards on the remains of the old field and as the process is continued the field goes travelling off into space with the speed of light, 3×10^8 *meters per second*, 186,000 *miles per second, or* 162,000 *nautical miles per second.*

All that has been said with reference to the magnetic field is equally true for the electric field, Fig. 10-1 (c). That is, the radiated field consists of magnetic and electric fields and is called an *electromagnetic* (*e-m*) *field.* The variation of magnetic or electric field strength with distance in opposite directions from the aerial is sinusoidal as shown in Fig. 10-1 (d). It should be carefully noted that the electric field is propagated in the vertical plane while the magnetic field is propagated in the horizontal plane. Another important property of the electromagnetic field is that the radiated electric and magnetic fields are in phase.

Within a few wavelengths from the transmitting aerial both the lines of magnetic and electric force are appreciably curved, but at great distances the curvature becomes so slight that the front presented to the receiving aerial by these lines becomes almost a vertical plane, as shown in Fig. 10-2. This plane is called the *wave front.*

10.3 Polarization. If one end of a rope that is stretched out and made

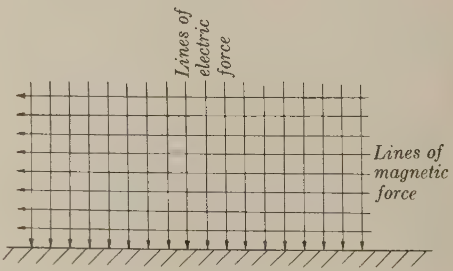


FIG. 10-2 Wave front of electromagnetic wave.

secure at the other end is given a violent up and down movement, a series of waves will travel along the rope, becoming rapidly attenuated as they reach the fastened end. The direction of movement of the rope is in a vertical plane through the direction of propagation and so these waves are referred to as being *vertically polarized*. The waves of the sea are vertically polarized.

If the movement of the end of the rope is horizontal, the waves are *horizontally polarized*. Similarly, if the movement of the rope is in a straight line making an angle with the vertical, the waves are *obliquely polarized*. Also, if the rope is given a rapid rotary movement of its free end a *circularly polarized* wave is produced. The wave consists of a vertical component and a horizontal component. If either of these components is made slightly larger in amplitude than the other an *elliptically polarized* wave is generated.

In all the above cases, however, the wave motion takes place in one plane only, through the direction of propagation and hence these kinds of waves are all referred to as *plane polarized waves*. Fig. 10-2 shows the wave front of a vertically polarized wave and Fig. 10-3 shows that of an obliquely polarized wave.

An electromagnetic wave is said to be vertically polarized if its electric field is in the vertical plane through the direction of propagation. Although there is a horizontal magnetic component, polarization of a wave is decided by the direction of its electric component.

It should be noted that no radio wave can travel far over a conducting surface such as the earth if it is horizontally polarized. If the wave is horizontally polarized the conductivity of the earth tends to short circuit any electrostatic component parallel to its surface, and further, if the magnetic field of the wave cuts the earth's surface circulating currents are formed to dissipate this component. For all radio-frequency waves up to approximately 25 Mc/s the earth acts as a good conductor and thus within this frequency range a vertical aerial is used to radiate a vertically polarized wave. As the frequency of the radio wave increases the conductivity of the earth decreases and a horizontally polarized wave becomes practical and so a horizontal aerial finds definite application at the higher frequencies.

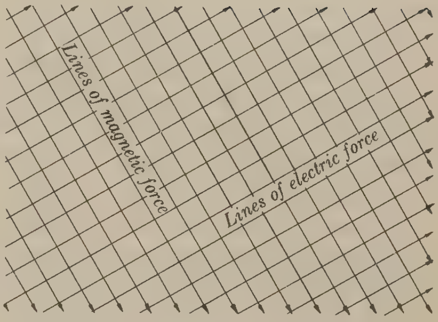


FIG. 10-3 Wave front of an obliquely polarized wave.

10.4 Definition of a transmission line. A *transmission line* is an arrangement to efficiently transfer energy from one point to another. In the early days of radio and relatively low operating frequencies, the technician had very little interest in transmission lines. However, the rapid growth of radio to the higher frequencies has made a thorough study of the principles and applications of transmission lines essential. Transmission lines have a wide application in modern radio and radar circuits. One of the most important applications of a transmission line is to transfer energy from a transmitter to an aerial so that maximum transfer results. A second important application is its use in feeding a complex aerial array in which various elements of the array must be fed in phase or out of phase with each other. A third application at the ultra-high-frequencies makes use of transmission lines as oscillator circuit elements.

A transmission line may be designed to conduct electrical energy at any given frequency but the considerations that follow are concerned in particular with the transfer of radio-frequency energy.

All transmission lines have two ends, the *input end* (*source end*) to which power is supplied and the *output end* (*load end*) to which the power is transferred. Thus, in a transmitting system, the input end is connected to the transmitter and the output end is connected to the aerial which becomes the load. In a receiving system the input end of the line is connected to the aerial and output end to the receiver.

10.5 Wave motion on a line. Since the basic application of an r-f transmission line is to efficiently transfer energy from a transmitter to an aerial so that maximum radiation results, the problems peculiar to this application are considered first.

It can be taken as a general rule that radiation from a conductor is greatest at points where the current is greatest. Thus, to obtain maximum radiation from an aerial the problem resolves itself into obtaining relatively large current values in the aerial. This is accomplished by producing in the aerial what are known as *standing waves* of current. An understanding of standing waves on an aerial is dependent upon an understanding of standing waves on a transmission line. Such understanding is facilitated by a study of the wave motion on an imaginary line of infinite length. An infinite line, while impossible to construct, provides an excellent subject for study because the behaviour on an actual line can be understood from that on the infinite line.

A. The Infinite Line

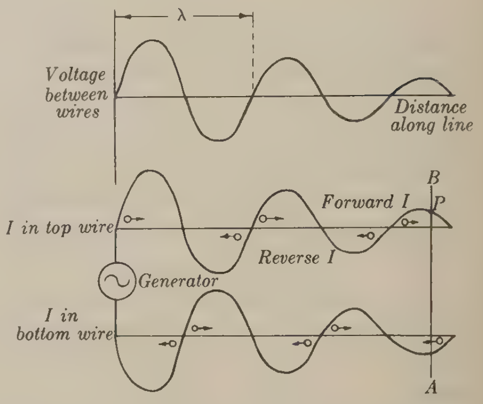


Fig.10-4 Voltage and current distribution on a two-wire line at one instant.

Fig. 10-4 shows the sine waves of current and voltage that exist on an imaginary two-wire line of infinite length when a generator is connected across it as shown. This figure illustrates the condition existing if the voltage and current are stopped for an instant in time just at the time when the alternating voltage of the source has reached zero. The voltage maxima and current maxima appear at the same points on this line, but because of line losses the values of current and voltage keep diminishing in amplitude down the line. Thus, the voltage and current are in phase and the operation can be pictured as a sine wave moving along the line or conductor at a definite velocity (v). As the sine wave moves past a particular point represented by the line AB, the value of the current at any instant is represented by the distance of point P from the horizontal axis.

Waves exist because it takes a certain amount of time for electrons to transfer energy down a wire by means of their motion. This time is negligible in electrical circuits used in commercial power work. For example, Fig. 10-5 shows the voltage wave of a 60-cycle generator. The wavelength of this 60-cycle voltage is 3100 miles and the voltage reverses polarity only at each

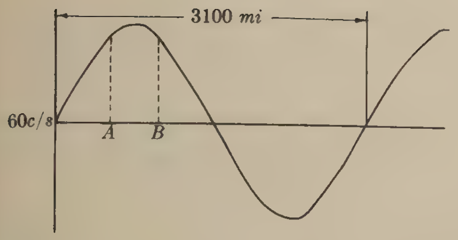


FIG. 10-5 Voltage wave of a 60-cycle generator.

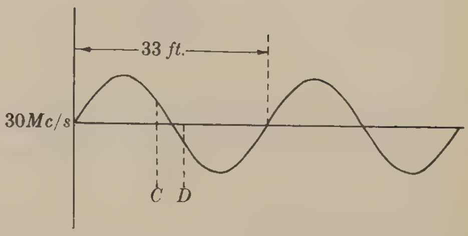


FIG. 10-6 Voltage wave of a 30 Mc/s generator.

half wavelength, that is, at 1550-mile intervals. Since the length of the normal power line circuits is usually less than 1550 miles no current reversals occur along the line at any one instant. The distance from *A* to *B* might represent 500 miles and since such a small part of the wave is being considered the voltage and current values at the input of the circuits are substantially the same as those at the output end and have the same polarity at the same instant.

However, if an alternator is producing a 30 Mc/s voltage as in an h-f transmitter, a positive voltage or current peak travels down the conductor only about 33 feet before another positive peak is put into the end of the line by the alternator as shown in Fig. 10-6. Thus the current in a wire is reversed completely within a few feet, as between *C* and *D*, at any instant and hence at high frequencies this effect must be taken into consideration.

It is noted that the phase relationship is the same at any two points along the current wave separated by any whole multiple of a wavelength (λ). Also there is a phase difference of 180° at any two points separated by a distance of $\frac{\lambda}{2}$ or an odd multiple of $\frac{\lambda}{2}$.

If the line continues uniformly the current and voltage always are in phase throughout its length but their amplitudes gradually decrease with distance from the generator. Energy supplied by the generator to the line never reaches the infinitely-distant load end. Consequently the energy is wasted within the line.

B. Open-Ended Line

1. REFLECTION AND STANDING WAVES.

If a line has a finite length and has an open end (Fig. 10-7) the impedance at the end can be considered as infinite. When energy is applied to the input end, waves of current and voltage sweep down the line in phase with each other. The initial current and voltage waves are in phase because the characteristics of the line at the input end appear the same as for an infinite line as the wave is starting toward the output end. The *in-phase* condition of these waves can be changed *only when they encounter a difference in the impedance between the two wires of the line.* When the wave of current reaches the open end of the line the current collapses to zero since there is no place for electrons to continue to move at this point. When the current collapses the magnetic field set up by it collapses. Consequently the collapsing magnetic field cuts the conductors near the output end and induces additional voltage across the line in the same direction. This voltage acts like a reverse generator and sets up new current and voltage waves which travel back on the line toward the input. These waves are referred to as *reflected* waves of current and voltage (Fig. 10-7).

It can be seen (Fig. 10-7) that the

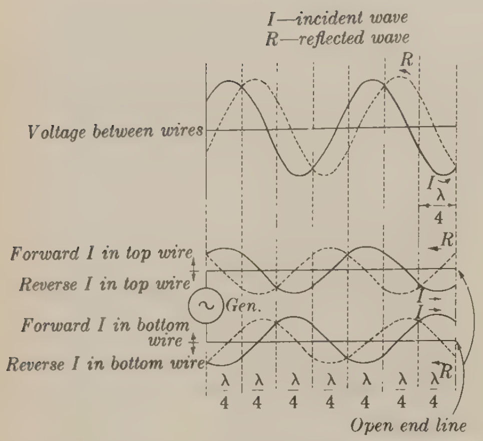


Fig. 10-7 Incident and reflected waves along an open-ended line at some instant.

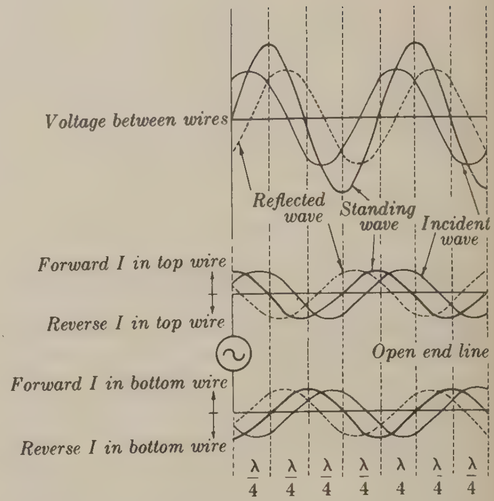


Fig. 10-8 Standing waves of current and voltage along an open-ended line at some instant.

beginning of the reflected wave of voltage starting back has the same polarity as the incident wave; thus the voltages at the open end add to produce a voltage which is the sum of the two. It is seen that the incident wave would reach zero a short distance beyond the end of the line if it were not reflected. Hence the reflected wave reaches zero the same short distance from the end of the line but in a direction toward the generator. At a distance $\frac{\lambda}{4}$ from the open end the forward-moving incident voltage wave and the backward-moving reflected wave will be in anti-phase since the total distance of $2 \times \frac{\lambda}{4}$ produces a phase difference of 180°. For a similar reason at a distance of $\frac{\lambda}{2}$ from the open end, the incident and reflected waves of voltage are in phase. If these voltages are added algebraically the resultant (Fig. 10-8) is called a *standing wave of voltage* and indicates the voltage condition that exists between the wires for the instant depicted in the diagram, that is, at an instant when the alternating voltage of the source has decreased slightly from its maximum value.

Since the total current at the end of the line is zero, the reflected current wave must be 180° out of phase with the incident current wave (Fig. 10-7). Thus the reflected current reverses in direction at the open end and the incident current and reflected current add to produce zero current. At a distance of $\frac{\lambda}{4}$ from the open end the forward-moving incident wave of current and the backward-moving reflected wave are in phase with each other since the total distance $2 \times \frac{\lambda}{4}$ produces an additional phase shift of 180°. Further, at a distance of $\frac{\lambda}{2}$ from the open end the incident and reflected waves are 180° out of phase. If these currents are added algebraically the resultant (Fig. 10-8) is called a *standing wave of current*. Again, this diagram indicates the current condition existing at the instant depicted.

A true picture of the standing-wave conditions existing may be obtained by a consideration of the conditions existing at various instants as the alternating voltage of the source decreases from a

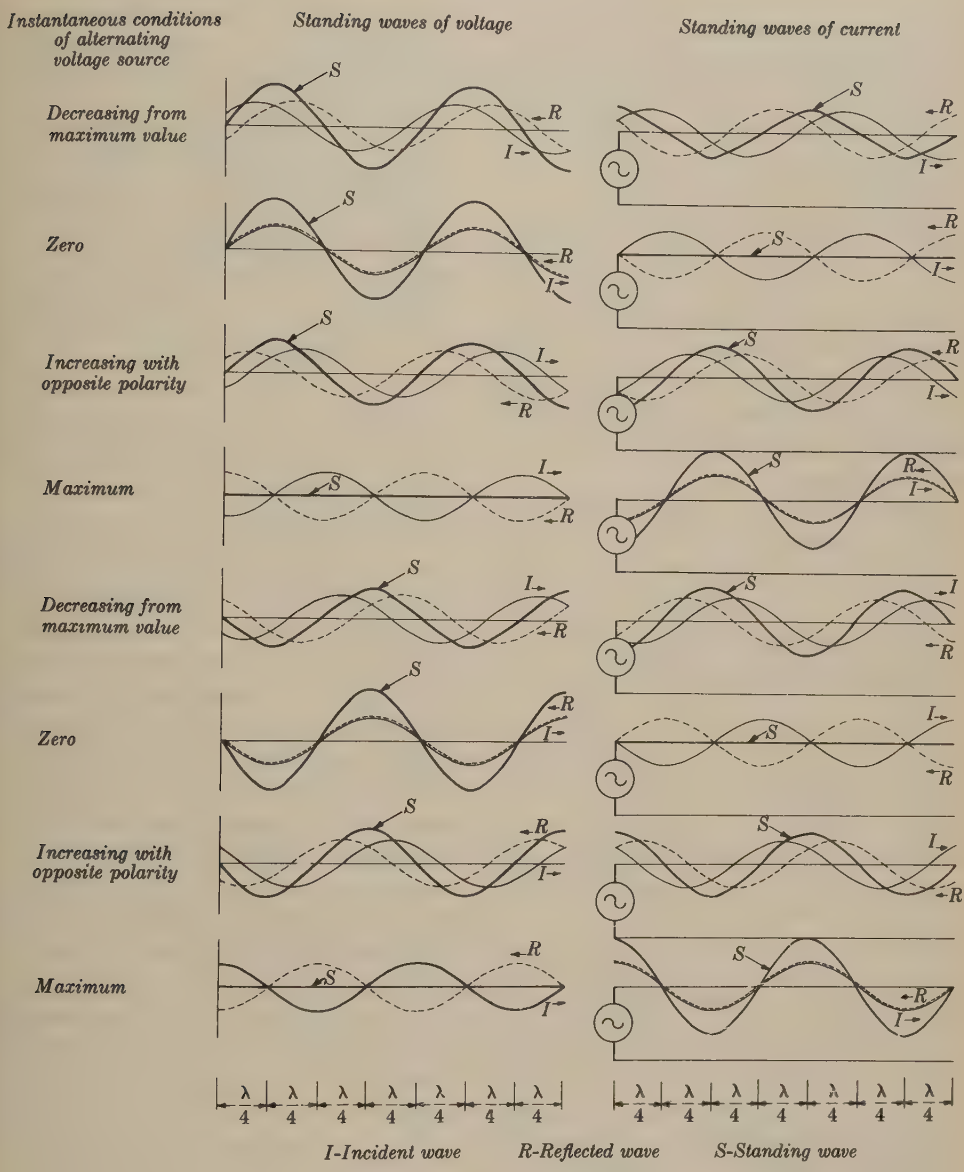

Instantaneous conditions of alternating voltage source
Standing waves of voltage
Standing waves of current
Decreasing from maximum value
Zero
Increasing with opposite polarity
Maximum
Decreasing from maximum value
Zero
Increasing with opposite polarity
Maximum
S
R
I
λ/4
I-Incident wave
R-Reflected wave
S-Standing wave

FIG. 10-9 Standing waves of current and voltage along an open-ended line at several instants during a complete cycle.

maximum to zero, reverses polarity and increases to a maximum and then decreases to zero again reversing polarity and increasing to a maximum to complete the cycle. Fig. 10-9 shows each of these instants with the incident waves, reflected waves and standing waves plotted. It can be seen that the incident and reflected waves of voltage at the output end always add to produce an alternating voltage which is the sum of the two. Therefore at the open end of the line or at any other point on the line there is a constant amplitude of *alternating* voltage which is the sum of the incident and reflected voltages. Further, for each different point along the line this amplitude is different and actually the standing wave is not stationary since the voltage at any one point is *alternating* and only the amplitude of the alternation at that point remains the same.

Standing waves of voltage can be measured by an r-f voltmeter and such a meter indicates a series of peak (or r.m.s.) values along the line which, when plotted, produce a standing wave of values. Points along the line at which voltage minima occur are called voltage *nodes* and those points at which voltage maxima occur are called voltage *antinodes* (or *loops*).

Current conditions on the line at various instants are similar to voltage conditions. At the open end the current is zero and hence the standing waves of current and voltage are 90° out of phase. At a distance $\frac{\lambda}{4}$ from the open end there is a current maximum (or a *current loop*) and at a distance of $\frac{\lambda}{2}$ from the open end there is a current node. Thereafter at each point an odd multiple of $\frac{\lambda}{4}$ from the open end there will be a current maximum and at each point an even multiple of $\frac{\lambda}{4}$ from the open end there will be a current minimum.

When an incident wave, that is, a *travelling wave* moves along a transmission line, the peak value of the wave moves along the line from point to point arriving at *different points at different times.* The *peak* value is the *same* for every point on the line, at the instant that the peak value is passing through that point. However this is *not* the case in a standing wave. In a standing wave every point along the line reaches its peak value at the *same instant* that every other point reaches its *peak* value. The *peak* value reached at any point on the line is *not* the value reached at adjacent points. Thus the peak value of current at one point on the line may be 1 ampere while at another point in the same half cycle the peak value may be 20 amperes, but both points reach their peak at the same instant of time. The location of the nodes and loops of a standing wave does not vary in a horizontal direction along the line but the instantaneous values of current at these points varies sinusoidally if the generated voltage is sinusoidal. Then, there is always a sinusoidal variation between the peaks of a standing wave. This can be understood from Fig. 10-10 which shows nine different instants in the same standing wave. On these successive instants the wave appears as shown by positions 1, 2, 3, 4, 5, 6, 7, 8, 9, 8, 7, 6, 5, 4, 3, 2, 1. Usually a standing wave is only shown in one of these positions.

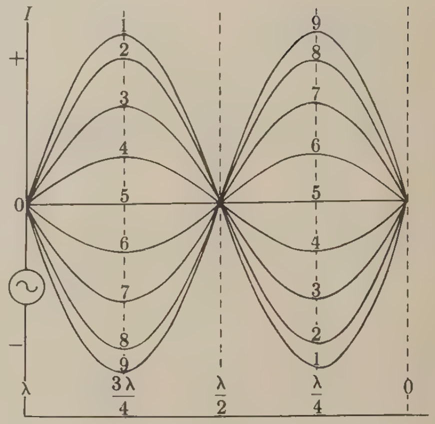


Fig.10-10 Appearance of a standing wave at nine instants during a complete cycle.

As the magnetic field about the wire is proportional to the current a field strength meter of the absorption type may be used to give an indication of the current strength along the line. It should be noted that for convenience in drawing only the current in one of the wires is shown.

2. Standing-Wave Ratios.

An indication of the amplitude of the standing wave may be obtained by comparing the voltage at a maximum position to that at a minimum position. The ratio showing this comparison is known as the *standing-wave ratio* and therefore may be expressed:

$$\text{Standing-Wave Ratio} = \frac{\text{max. voltage}}{\text{min. voltage}}$$

Thus if an r-f voltmeter is used and the voltage between the wires at a minimum point is 5 and the voltage between the wires at a maximum point is 15, then

$$\text{Standing-wave ratio} = \frac{15}{5} = \frac{3}{1} = 3{:}1$$

3. MEASUREMENT OF WAVELENGTH USING STANDING WAVES.

The distance between two successive maxima or two successive minima is equal to one half wavelength at the applied frequency. This distance may be measured and used to determine the wavelength and thus the frequency. This is assuming that the velocity of propagation along the wires is the same as that through free space or air which is not true, but for most practical work it is considered close enough to be acceptable. In effect the velocity of propagation is different for each different medium and hence is different for different conductors. Thus measurements of wavelengths would be different for different conductors and so the *apparent wavelength*, for accurate results, must be corrected to give the *true wavelength* thus:

In air

$$\lambda = \frac{\mathbf{984}}{\textbf{frequency in megacycles}}$$

In any medium

$$\lambda = \frac{\mathbf{984k}}{\textbf{frequency in megacycles}}$$

where λ = wavelength in feet

and k = a constant depending on the type of line. This is the factor which expresses the ratio of the actual velocity of the energy on the line to the velocity of light. The following are some typical values of k:

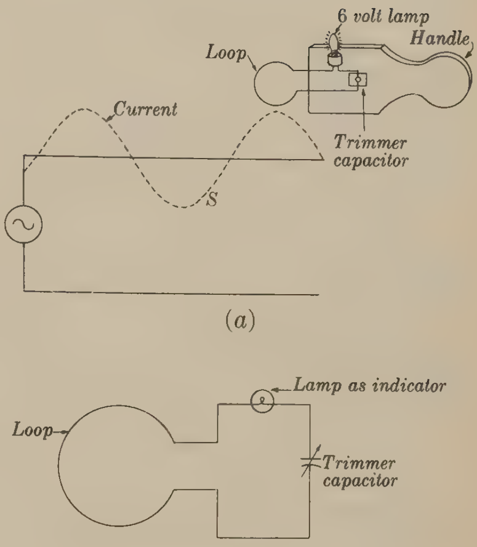


FIG. 10-11 Absorption type standing-wave indicator (a) Position of indicator for maximum indication (b) Equivalent circuit of indicator.

Two-wire parallel conductor	k = 0.975
Two-conductor parallel tubing	k = 0.950
Concentric line (air insulation)	k = 0.850
Concentric line (rubber insulation)	k = 0.560 to 0.650
Twisted pair	k = 0.560 to 0.650

4. STANDING-WAVE INDICATORS.

Standing waves of current or voltage can be detected by means of an absorption type indicator consisting of a resonant circuit (Fig. 10-11) employing a single loop of wire and a small variable capacitor of the trimmer type in series with a neon or 6-volt dial light. If this indicator, tuned to the frequency of the generator, is moved along the length of the wire and slightly above it the lamp glows brightly opposite a current maximum and dimly opposite a current minimum. As this device is not a sensitive one the exact point at which a current loop or node occurs cannot be

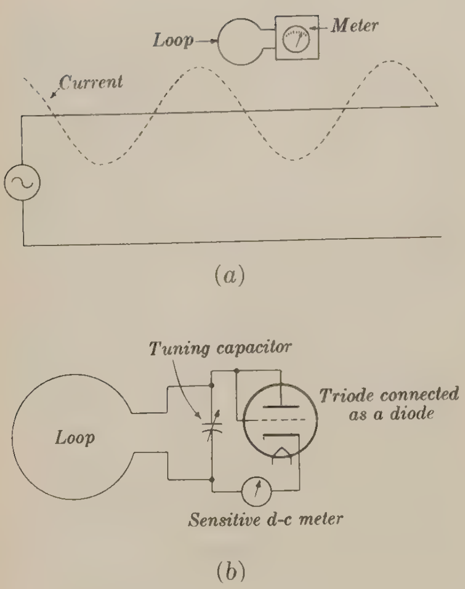


FIG. 10-12 A more accurate standing-wave indicator (a) Position of indicator for maximum indication (b) Equivalent circuit of indicator.

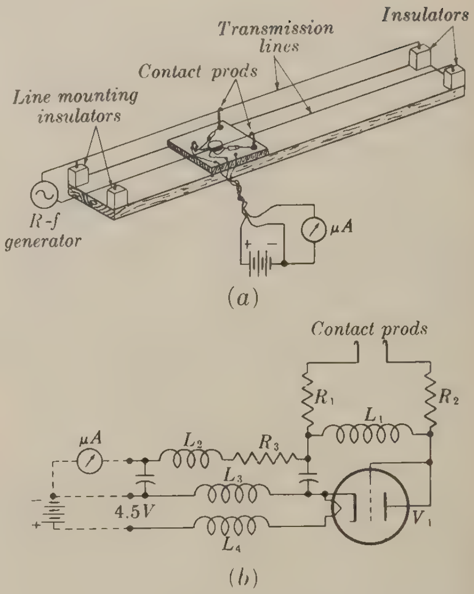


FIG. 10-13 A voltage standing-wave indicator (a) Position of indicator relative to line (b) Equivalent circuit of indicator.

accurately determined. If these results are used to calculate wavelength and frequency only approximate values are obtained.

Further, if the loop is placed across and touching the wires of the line, then moved along their length, standing waves of voltage will be indicated. Again only approximate results are obtainable.

More accurate results may be obtained by modifying the indicator as in Fig. 10-12. The triode connected as a diode is used as a rectifier so that a sensitive d-c meter may be used and operated by the r-f energy picked up by the loop. As the meter is connected across the tuned circuit it will read a maximum when the loop is opposite a current maximum and a minimum when opposite a current minimum. As meter readings at maximum and minimum points may be obtained and compared standing-wave ratios may be determined.

In most cases more convenient and more accurate measurements are obtained if an r-f voltage indicator is used to locate and measure voltage maxima and minima as a less sensitive meter may be used. A practical untuned arrangement is shown in Fig. 10-13. This vacuum-tube voltmeter employs an acorn high-frequency triode V_1 connected as a diode and is mounted breadboard style to allow it to be moved along beneath the transmission line. Two short prods of copper wire make contact with the wires of the line and the voltmeter is isolated by the resistors R_1 and R_2 of 1 megohm each so as not to load the transmission line. The voltage between the wires is applied across the voltage divider R_1, R_2 and L_1. The L_1 voltage is applied to V_1 and an indication of its value is give by the reading on the microammeter in the cathode circuit of the valve. A 50,000 ohm resistor, R_3, in series with a 2.5mH choke L_2, is used as the diode load. Three dry cells connected in series are used to provide filament supply to the valve and two inductances L_3 and L_4 each of 20 turns of number 20 wire wound on $\frac{5}{8}''$ diameter

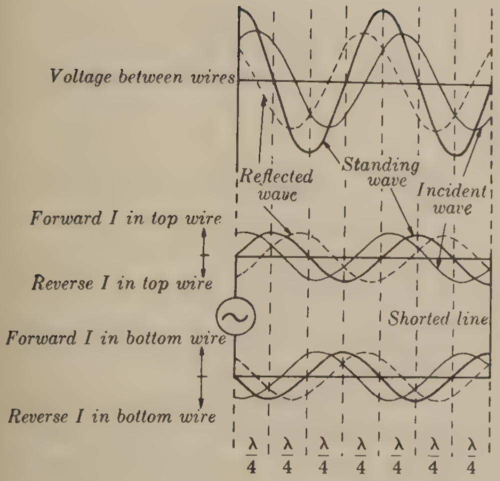


FIG. 10-14 Standing waves of current and voltage along a shorted line at some instant.

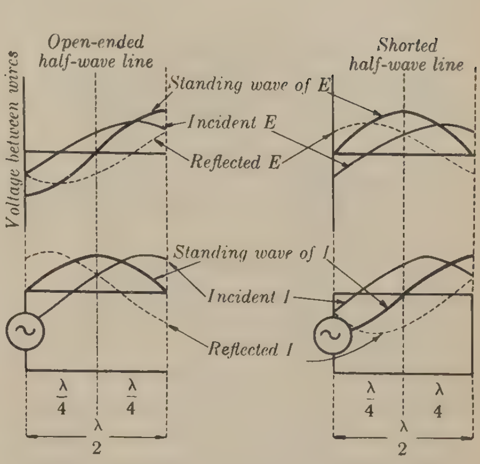


FIG. 10-15 Standing waves of current and voltage along an open and a shorted half-wave line at some instant.

formers, are used to isolate the filament supply from the r-f. As the meter is moved to a position of maximum voltage on the line, a maximum current reading is indicated on the microammeter.

C. The Shorted Line

If an r-f generator is connected across a transmission line whose output end is short-circuited (Fig. 10-14) the conditions that develop in this line are very similar to those that develop in an open-ended line. A travelling wave sweeps down the line to the short-circuited output end and since the short circuit is a condition of zero impedance the energy cannot be absorbed and reflection takes place. The current at the shorted end of the line is a maximum and the voltage is a minimum. The reflected waves and the incident waves combine along the line and standing waves of current and voltage are set up. A comparison of Fig. 10-14 with Fig. 10-8 shows that the voltage and current relationships are exactly opposite in a short-circuited line to what they are in an open-ended line. The points of maximum and minimum current and voltage are determined from the output end of the line where reflection begins. A line does not have to be any particular length to produce standing waves; the only requirement is that reflections must occur, that is, all the energy of the incident wave must not be absorbed at the output end. The standing waves of current and voltage at any instant which appear on an open half-wave line and a shorted half-wave line are shown in Fig. 10-15.

10.6 Characteristic impedance of a transmission line. Long uniform transmission lines presented a problem to early communications engineers because of the fact that the electrical constants of such lines are distributed in space rather than lumped at one point. Hence the engineering and design problems could not be dealt with by familiar circuit analysis methods and new methods have been developed.

At low frequencies it is simplest to think of a current flowing out one wire of the transmission line and back in the other. Thus the product of the current and the potential difference at the sending end gives the power delivered to the line. Also, the product of the current and the potential difference at the output

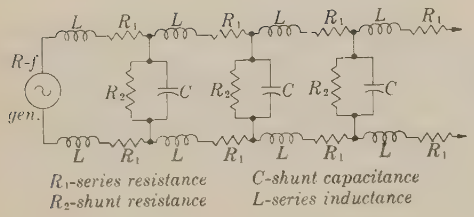


FIG.10-16 Equivalent circuit of a two-wire line in terms of electrical constants.

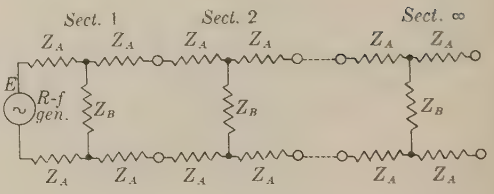


FIG.10-17 Equivalent circuit of a two-wire line in terms of lumped constants.

end gives the power delivered to the load. At radio frequencies where one wavelength is only a fraction of the length of the transmission line the above conception is not applicable because current flows in opposite directions in the same conductor in each successive half wavelength. Further, there is also an impedance to any flow of current and to any change in the rate of flow and there is dissipation of energy called *attenuation*. For simplicity it is best to think of the transmission line as having many distributed electrical constants which form a network and which can be resolved into some total effective impedance in order that the effect of the line on the r-f wave-trains may be determined.

Any form of two-wire transmission line has four fundamental electrical constants or parameters:

(i) series resistance
(ii) shunt resistance
(iii) series inductance
(iv) shunt capacitance.

Thus any line possesses resistance that limits the current in the line. In addition, in a transmission line it is impossible to achieve perfect insulation at r-f between the wires of the line so that there is a certain amount of leakage or shunt resistance between the wires. Furthermore, since a magnetic field is set up about any current-carrying conductor the line possesses series inductance. The two wires of the line act as small capacitor plates with each other and with ground so the line has shunt capacitance. These electrical constants are illustrated in Fig. 10-16.

For convenience in understanding, all of these various effects of resistance, capacitance and inductance can be lumped together as impedances which act both in series with and across the line. The line can be considered as made up of a great many similar sections each with a definite amount of series and shunt impedance. This is shown in Fig. 10-17 which represents unit sections of a line which stretches out to infinity.

If all sections except the first are disconnected the current flowing into the line will be very small and is given by

$$I = \frac{E}{2Z_A + Z_B}$$

If a second section is added to the first there is a greater current flow, since Z_B is now shunted by an impedance equal to $4Z_A + Z_B$. This process can be continued by adding more sections to the line, each of which causes an increase in the current over that caused by adding the preceding section. Eventually, the point is reached where the addition of another section does not have appreciable effect upon the current. This is illustrated in Fig. 10-18 which shows how the current tapers off to a definite value as the number of unit sections, and hence the length of the line, is increased. Therefore, it may be concluded that a transmission line of infinite length assumes one definite value of impedance. This value is known as the *characteristic impedance*, Z_O, of the line. Different

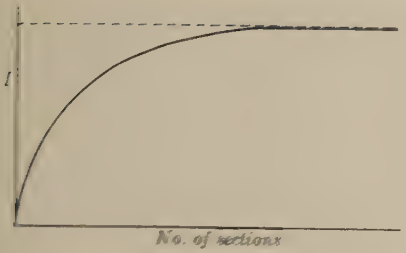


Fig. 10-18 Variation in current amplitude with number of sections of a line.

types of transmission lines have different characteristic impedances.

Fig. 10-19 is a numerical example which shows how small sections of the line can be added to obtain the characteristic impedance. Consider a unit section to have an effective resistance made up of a series resistance of 10Ω and a resistance of 100Ω across the line. Thus the impedance across the generator end of line, Z_{XY}, is 110Ω. In Fig. 10-19 (b) a second identical unit section is added and the impedance Z_{XY}, 62.4Ω, is calculated from the rules for combining series-parallel networks. The process of adding one section is continued in each diagram from (c) to (h) and the impedance across XY is calculated and recorded on the diagram. Each time a section is added the impedance is decreased and it continues to decrease to the value 37.0Ω. Several hundred more sections can be added without any noticeable reduction in the impedance Z_{XY}. Thus the value 37.0Ω can be considered to be the characteristic impedance Z_o of the line.

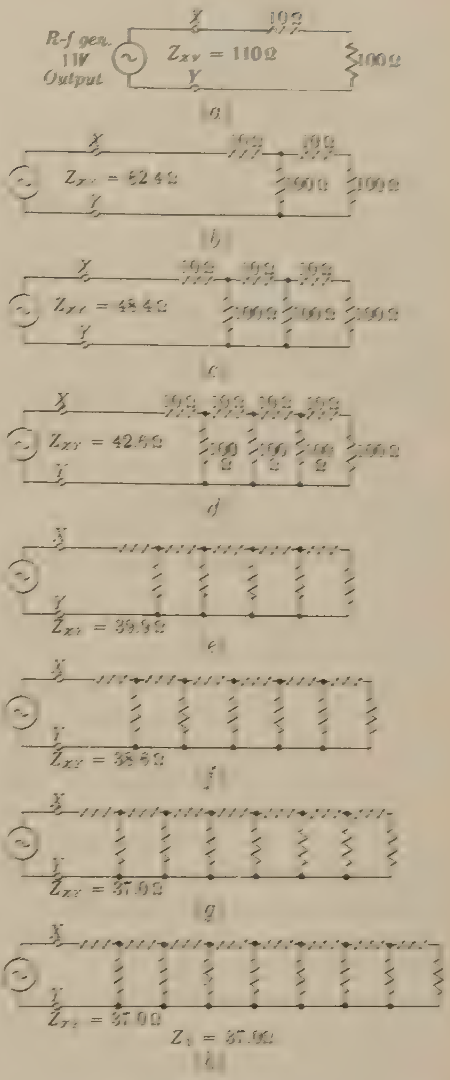


Fig. 10-19 Characteristic impedance of a line obtained by adding unit sections, one at a time.

The characteristic impedance of an infinite line has a very valuable use. If a load equal to this impedance Z_o is placed across the output end of any short or convenient length of the line the same effective impedance Z_o appears at the input terminals of the line. Only one value of impedance, the characteristic impedance, for any particular type and size of line acts in this way. This important fact can be verified by a continuation of the numerical example of Fig. 10-19. If the line of Fig. 10-19 (a) is terminated with its characteristic impedance Z_o, 37.0Ω, as in Fig. 10-20 (a) and the effective impedance calculated it is found to be 37.0Ω also. Thus if R is the impedance of Z_o and the 100Ω in parallel,

$$\frac{1}{R} = \frac{1}{37.0} + \frac{1}{100}$$

$$\therefore \frac{1}{R} = \frac{137}{3700}$$

Therefore, $R = \dfrac{3700}{137} = 27.0\Omega$

Hence the effective impedance is

$$Z_{XY} = 27.0+10 = 37.0\Omega$$

Similarly if a line consisting of three sections as in Fig. 10-20 (b) is terminated with its characteristic impedance Z_O, 37.0Ω, the overall effective impedance across XY can also, by calculation, be shown to be 37.0Ω.

Therefore there is one value of impedance, Z_O, which the line assumes if it is very long and, further, if this value of impedance is used to terminate a shorter line, the same effective value of impedance appears at the input.

The ability to determine the characteristic impedance of a transmission line provides a method for ensuring that maximum energy is transferred from generator to line to load. For maximum transfer of energy the impedance of the load must be the same as the impedance of the generator. The impedance seen by a generator looking into an infinite transmission line or into a line terminated in its characteristic impedance is simply the characteristic impedance of the line. Therefore to send maximum energy into the line it is only necessary to construct a transmission line with a characteristic impedance equal to the impedance of the generator, or otherwise it is necessary to match the impedance of the line by a transformer or other matching device to the impedance of the generator.

If maximum energy is to be transferred from the line to the load the impedance of the load must be matched to the impedance of the line. If the line is terminated with a load that has an impedance equal to the characteristic impedance of the line this condition is fulfilled. In addition, terminating the line with a load equal to the Z_O of the line does not change the effective load on the generator as the impedance that the generator sees still remains Z_O. However, if a line is connected to a load which has an impedance different from the Z_O of the line a wave reflection is produced back along the line as all the energy is not absorbed by the load. Thus standing waves are set up in the line and I^2R and radiation losses are increased in the line due to the increase in current. In this way maximum energy is not dissipated in the load.

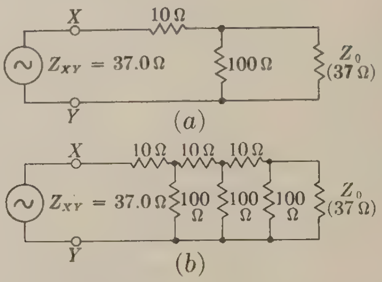


FIG. 10-20 Impedance of any length of line terminated by its characteristic impedance (a) One-section line (b) Three-section line.

The characteristic impedance of a two-wire line depends upon the size and spacing of the conductors and the kind of dielectric medium which separates the conductors. For a pair of parallel conductors, the characteristic impedance is given by the formula,

$$Z_O = k \log \frac{b}{a}$$

where Z_O = characteristic impedance
b = distance between the centres of the two conductors
a = radius of conductors
k = constant depending on the dielectric medium between the conductors.

From this formula it can be seen that the characteristic impedance decreases as the spacing is made smaller and the diameter of the conductors is made larger.

The construction of the common types of transmission lines together with calculation of their characteristic impedances are discussed in section 10-12.

10.7 Aerial impedance. In radio communications and radar, transmission lines are used to transfer r-f energy from the transmitter to the aerial. It is essential that the aerial which forms the load be matched to the generator (the transmitter) for maximum transfer of energy. Thus it is necessary to understand the meaning of aerial impedance so that a match may be effected.

First consider an open-ended transmission line with each wire one quarter wavelength long connected across an r-f generator as shown in Fig. 10-21. The standing waves of current set up on the wires of this line as a result of the reflection from the open end are shown in Fig. 10-21 (b) and the standing wave of voltage between the wires is shown in Fig. 10-21 (a). If the wires of the line are now spread as shown in Fig. 10-21 (c) a half-wavelength aerial usually called a *dipole* is formed and of course the standing waves of current and voltage on this basic aerial remain the same. The distribution of current and voltage on a half-wavelength aerial is shown in Fig. 10-21 (c). The voltage maxima appear at the ends with minimum voltage at the centre of the aerial, which is also the point of maximum current. The spreading of the quarter-wave sections of the line causes the fields radiated to become additive whereas the fields radiated from two closely spaced wires of Fig. 10-21 (b) cancel since the current in, and the field radiated from, one is 180° out of phase with the current in, and the field radiated from, the other.

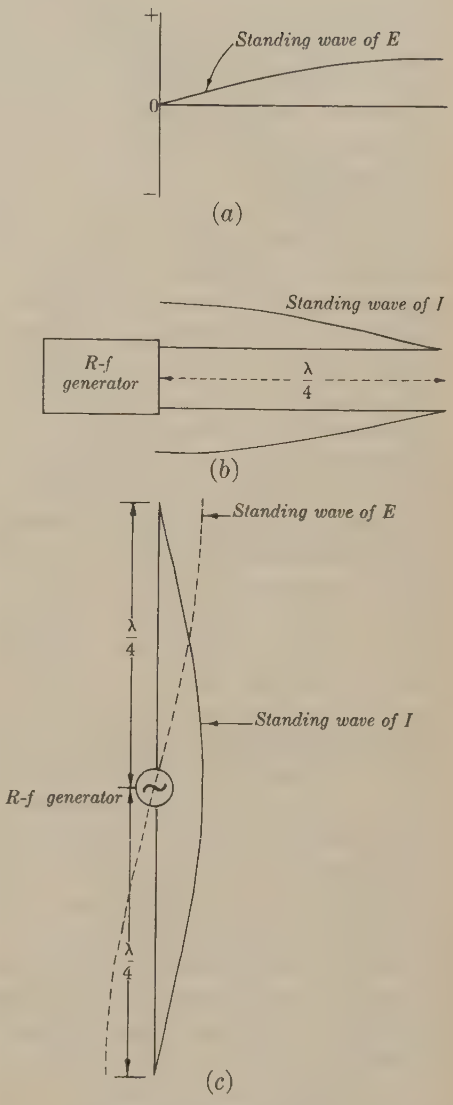


FIG. 10-21 Evolution of a dipole from an open half-wave two-wire transmission line (a) Standing wave of voltage (b) Standing wave of current (c) Standing waves on resultant dipole.

The impedance of an aerial is a quantity that varies from point to point along its length. Since the impedance can be considered equal in magnitude to the voltage divided by the current, that is $\frac{E}{I}$, a consideration of the current and voltage distribution on the aerial shows that the impedance of the aerial varies from a maximum at one end (that is, high E and low I), to a minimum at the centre (low E and high I), to an equal maximum at the other end.

At the centre of the aerial the voltage is a minimum and the current is a maximum. This corresponds to a capacitance and inductance in series at the resonant frequency. Therefore, at this centre point the voltage is in phase with the current, so the impedance at the centre point is resistive in character. At other points along the length of the half-wavelength aerial the impedance contains a reactive component. The resistance at the centre of any half-wave aerial has a fixed value approximately equal to 73Ω if the aerial is mounted a sufficient distance from such modifying influences as the surface of the earth. The centre point resistance, known as the *radiation resistance* of the aerial, is the equivalent of the physical resistance which would absorb an amount of energy from the transmitter equal to that radiated from the aerial.

A knowledge of the radiation resistance and the manner in which the impedance varies along the half-wave aerial is important in discussing the methods by which the output of a transmitter may be fed to the aerial.

10.8 Non-resonant transmission lines. In the preceding sections lines terminated in an open end and in a short circuit were discussed. In these cases it is found that reflection takes place at the end and standing waves are set up on the line. A transmission line on which there are no standing waves is called a *non-resonant line*. However, for a line to be considered non-resonant, it is not necessary for the line to have no standing waves on it but it is necessary that the standing-wave ratio be close to unity. The current and voltage waves along a non-resonant line are approximately in phase (Fig. 10-22); the line operates at a power factor of approximately unity and the most favourable conditions for maximum transfer of energy exist. A line should be made non-resonant to be used effectively to transfer r-f energy. A line can be made non-resonant by the proper termination of its output end.

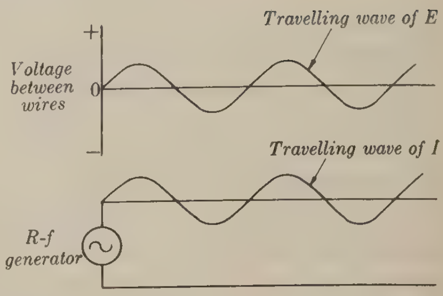


Fig.10-22 Voltage and current waves along a non-resonant line.

A non-resonant line is either an infinitely long line or a line terminated in its characteristic impedance. In each of these cases there are no reflections and all the energy sent down the line is absorbed at the output end. On lines carrying r-f currents the characteristic impedance is almost pure resistance. Therefore to make a line non-resonant it is terminated in a resistive load equal to the characteristic impedance of the line.

A non-resonant line, sometimes referred to as a *flat line* or an *untuned line* is desirable for the following reasons.

(i) The I^2R losses in the line are reduced when standing waves are eliminated because the high current peaks are not present.

(ii) There are no points of high r-f voltage, which is important when insulation must be provided where high power is used. Reduction in voltage also decreases dielectric losses in the line.

(iii) The two-wire non-resonant line consists of two parallel wires held apart by insulating spacers. The separation between the wires is kept small with respect to the wavelength under consideration. When the separation is small, that is, of the order of 0.01λ, there is very little radiation from the feeder lines. This is because the current in each wire at any point is 180° out of phase with the current in the other wire and so the radiation fields from the wires of the line almost cancel each other. The closer the wire spacing the more nearly do the radiation fields cancel. Thus the effective radiation from a two-wire line is small and as a result the radiation losses from the line are also small.

Since the resultant radiation from a two-wire line also depends upon the magnitude of the current flowing, a non-resonant line radiates less than one with standing waves present on it. For the same reason a non-resonant line picks up less energy when it is used to connect an aerial to its receiver. In the former case distortion of the directional radiation from a directional aerial is reduced and in the latter case there is a desirable increase in the signal-to-noise ratio.

When using a non-resonant line to connect a transmitter to its aerial, the system must be adjusted to prevent standing waves on the line. It has been seen that a transmission line terminated in its characteristic impedance transfers the maximum amount of energy to the termination. Therefore, if the transmission line is terminated in an aerial which provides a resistive load equal to the characteristic impedance of the line, correct conditions for maximum energy transfer are established. Several methods of matching an aerial to a non-resonant transmission line are considered in the following sections.

A. Delta Match to a Half-Wave Aerial

The impedance of a two-wire line is usually several hundred ohms so that the line cannot be connected directly to the centre of a half-wave aerial which has an approximate impedance at this point of 73 ohms. However, advantage can be taken of the fact that the aerial impedance increases as the distance increases from the centre point. This can be understood from a consideration of the current-voltage distribution on the aerial (Fig. 10-23). At the centre point of the aerial the voltage is a minimum and the current is a maximum, hence the impedance is $Z = \frac{E}{I} = \frac{0}{I} = 0.$

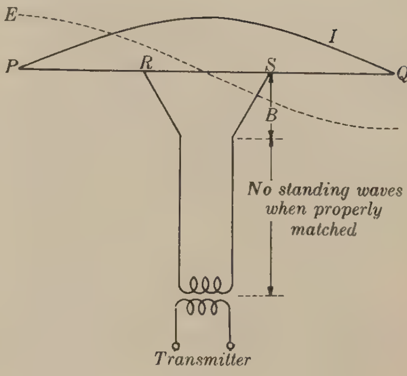


FIG. 10-23 Delta match to a half-wave aerial.

Also, voltages at P and Q are equal and opposite and hence the p.d. between P and Q is $2E$ while the current at P and Q (the open ends) is zero. Hence the impedance between P and Q is

$$Z = \frac{2E}{I} = \frac{2E}{0} = \infty.$$

Thus, if two points R and S are considered the voltages at R and S are equal and opposite and less than at P and Q while the current at R and S has increased from its zero value at P and Q. Therefore the impedance between R and S is intermediate in value between zero and infinity. Hence the transmission line may be connected to the dipole and the tapping points spread out or fanned out as shown in Fig. 10-23 until an aerial impedance is reached which just matches the impedance of the line. The match is indicated by the absence of standing waves on the line which in turn is indicated by means of a standing-wave indicator (Sec. 10.4). This method of effecting a match between an aerial and transmission line is called the *delta match*.

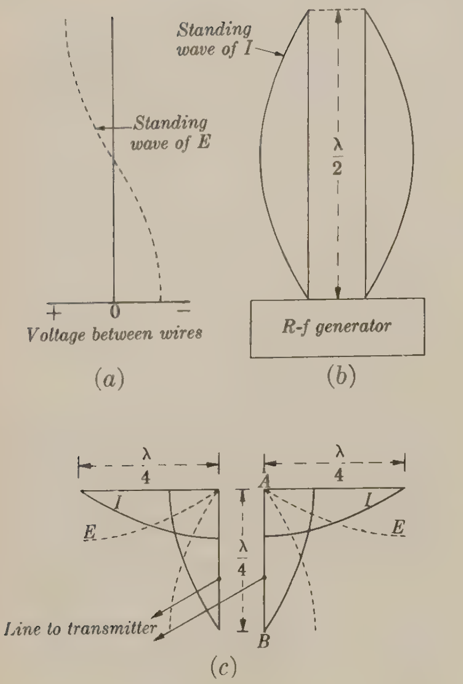


FIG. 10-24 Quarter-wave stub match to a dipole (a) Standing wave of voltage (b) Standing wave of current (c) Standing waves on stub and dipole.

B. Quarter- and Half-Wave Stub Matches

Another method of matching a transmission line to an aerial is by means of a *matching stub* which may be either one quarter wavelength or one half wavelength long.

Consider first the standing waves of current and voltage produced on a two-wire line which is one half wavelength long as shown in Figs. 10-15 and 10-24 (a) and (b). Now suppose that the half-wave line is turned back and through 90° at a point one quarter wavelength from its end as shown in Fig. 10-24 (c) to form a flat top half-wave aerial and a quarter-wave section or stub of line, extending down from the centre of the aerial, which is used to effect the match. The current-voltage distribution on the half-wave aerial and on the stub remains the same as before the line was folded and this distribution is indicated again in Fig. 10-24 (c). The quarter-wave stub has a voltage maximum at its end which is left open. As a result of the standing waves of voltage and current on the stub the impedance varies along its length. At the centre *A* the voltage is zero, the current is a maximum and thus the impedance is zero. Across the end of the stub *B* the voltage is a maximum and the current zero and hence the impedance is infinite. Thus a connecting point can be chosen to provide an aerial impedance which is just equal to the characteristic impedance of the line. Such a matching point can be found by moving the line tapping points along the stub until the point is reached where no standing waves are produced on the line.

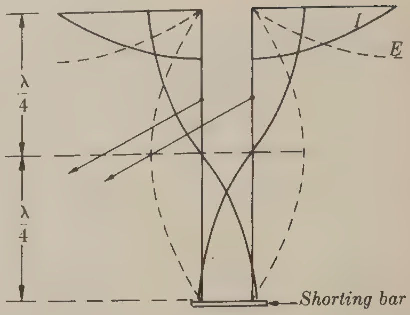


FIG. 10-25 Half-wave stub match to a dipole.

Similarly a half-wave stub may be used as shown in Fig. 10-25 but in this case to provide correct current voltage distribution there must be a voltage minimum at the end of the stub and so it is shorted by means of a conducting bar. The standing-wave distribution on the stub and on the half-wave aerial is shown in the diagram and it can be readily seen that the impedance varies from zero at the end of the stub, to a maximum at the $\frac{\lambda}{4}$ position from the end, to a minimum at its centre and then repeats. Thus there are two positions at which the transmission line can be connected to the stub so that the aerial impedance is matched to the characteristic impedance of the line.

10.9 Resonant lines. A *resonant transmission line* is a line having standing waves present on it. An open-ended line, shorted line, or a line terminated in a load which is not resistive or not equal to the characteristic impedance have standing waves on them and are therefore resonant lines. The separation between the successive peaks of the standing waves of either voltage or current is one half wavelength long. The distance representing a wavelength depends upon the frequency of the source. Thus a transmission line which is one half wavelength long for a given frequency is not one half wavelength long at another frequency. By varying the length of the line it can be tuned to the frequency of the generator, that is, tuned through resonance in a manner similar to the tuning of a conventional inductance-capacitance resonant circuit. Thus the resonant line is often referred to as a *tuned line.*

A resonant line has the following characteristics.

(i) It does not require an impedance match between line and load (aerial).

(ii) It provides some degree of frequency flexibility otherwise unobtainable.

(iii) It has high I^2R losses as a result of the high standing waves of current present on the line.

(iv) For the same reason its radiation losses are great.

A resonant-line may be used to feed an aerial but it finds its widest application in the use of resonant-line sections as impedance matching devices. A consideration of the resonant line as an aerial-feed method is considered first and a discussion of resonant-line sections follows.

A. Resonant-Line Centre Feed of an Aerial

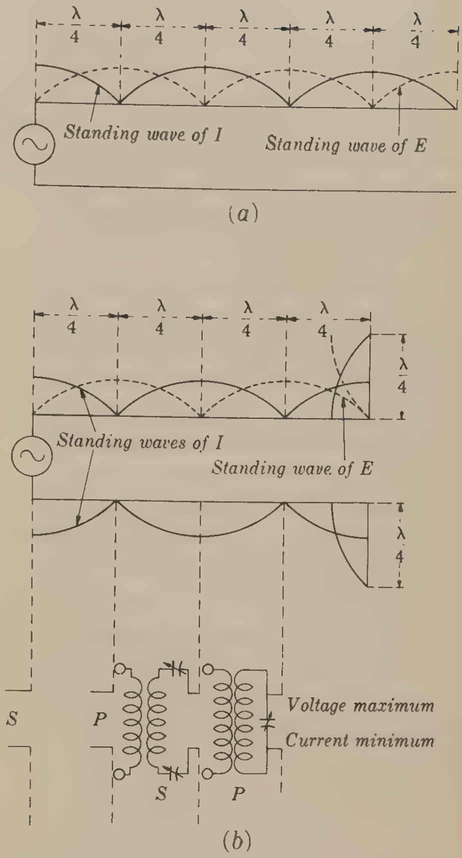


FIG. 10-26 Resonant line centre-feeding a dipole (a) Standing waves of current and voltage on line (b) Standing waves of current and voltage on line and dipole.

The method of feeding a half-wave aerial by means of a resonant line may be understood from a consideration of Fig. 10-26. Fig. 10-26 (a) shows an open-ended transmission line which is cut to an exact number of quarter wavelengths at a given frequency and the standing waves of current and voltage are indicated on the line. If each wire of the line is now folded back through 90° at a point $\frac{\lambda}{4}$ from the end as shown in Fig. 10-26 (b) a resonant line feeding a half-wave aerial results. The current-voltage distribution along the line and the aerial remain the same but now radiation from the aerial sections become additive while radiation from the wires of the line still cancel one another if the wires of the line are close together. At a point $\frac{\lambda}{4}$ from the aerial along the line there is a current minimum and a voltage maximum. If the transmitter is connected at this point the coupling must furnish maximum voltage and minimum current. This condition is satisfied by the parallel tuned coupling labelled *P* in the figure. The parallel-resonant type of coupling is used when connection is made to the transmission line at any point which is an odd number of quarter wavelengths from the aerial.

At points an even number of quarter wavelengths from the aerial occur current maxima and voltage minima. Thus if the transmitter is to be connected at these points the coupling required must furnish maximum current and minimum voltage. The series-resonant coupling shown at *S* in the figure meets these conditions.

B. Resonant-Line End Feed of an Aerial

The half-wave aerial may also be fed at one end by means of a tuned transmission line (Fig. 10-27). The current-voltage distribution on the line and aerial is indicated in the diagram. Although one end of the tuned line is unconnected in this system, the unbalance introduced thereby is small if the feeder length is an accurate multiple of the quarter wavelength. The transmitter coupling to the end-feed tuned line is the reverse of the centre-feed aerial with respect to the number of quarter wavelengths from the aerial. Since the current maxima now occur at points an odd number of quarter wavelengths from the aerial, the series-resonant coupling is used at these points. The voltage maxima and current minima are at points an even number of quarter wavelengths from the aerial, requiring parallel-resonant coupling.

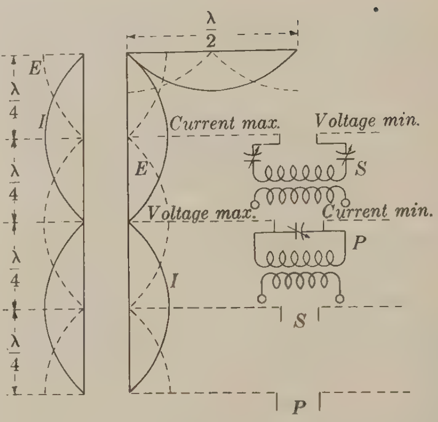


FIG. 10-27 Resonant line end-feeding a half-wave aerial.

The tuned lines are usually parallel wires held apart by insulating spacers. The separation between the wires must be kept small with respect to the wavelength under consideration. When the separation is small, perhaps of the order of 0.01λ, there is very little radiation from the feeder lines. The current in each wire at any point is 180° out of phase with the current in the other wire so that the radiation fields from the wires of the line effectively cancel each other.

C. Resonant Feed for Multiple-Frequency Operation

It sometimes is convenient and desirable to use the same half-wave aerial for operation on several frequencies. This can be done by the use of resonant-line feed. When only a half wave of current and voltage are present on the aerial the aerial is operating at its fundamental frequency or 1st harmonic (Fig. 10-28). If the aerial is operated at the second harmonic, that is twice the fundamental frequency, there is a full wavelength of current and voltage standing on the aerial. If the aerial is operated at the third harmonic, that is three times the

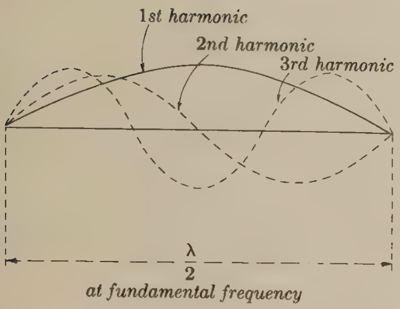


FIG.10-28 Operation of a half-wave aerial at its first, second and third harmonic frequency.

fundamental frequency, there are three half wavelengths of current and voltage standing on the aerial.

In order to permit a half-wave aerial to operate at its harmonic frequencies it is only necessary to use a resonant transmission line fed in the proper manner as shown in Fig. 10-29. Thus in Fig. 10-29 (a) the resonant line is current fed by means of a series circuit; in (b) the resonant line is voltage fed to permit operation at the 2nd harmonic; in (c) current feed provides satisfactory conditions for 3rd harmonic operation. The shift from current to voltage feed is easily accomplished in service transmitters by a control on the front panel marked to indicate the type of coupling circuit, series (current) or parallel (voltage).

10.10 Comparison of resonant-line sections with conventional resonant circuits. A resonant transmission line acts as a resonant circuit tuned to some particular frequency. This fact is important in the application of resonant lines as circuit elements at the higher frequencies and as a result a clear understanding of how a resonant line presents the characteristics of a conventional resonant circuit is necessary.

In order to act as a resonant circuit the line is either open-ended or short-circuited at the output end and it is cut to some multiple of a quarter wavelength.

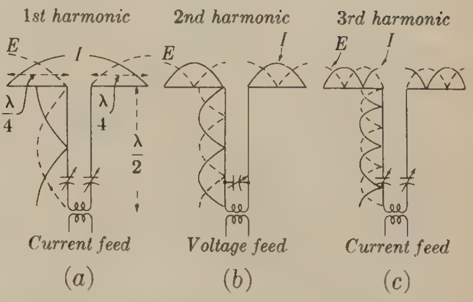


FIG.10-29 Types of feed to permit operation of a dipole at harmonic frequencies (a) First harmonic (b) Second harmonic (c) Third harmonic.

A. Open-Ended Resonant-Line Section

Fig.10-30 illustrates the relationship between voltage, current and impedance for various sections, or lengths, of open-ended resonant lines. The impedance which the generator sees is shown at the top of the chart. The impedance curve indicates the relative values of impedance presented to the generator as it is moved from right to left along the line. The impedance values are derived from the very simple principle that the impedance at any point is equal to the voltage at that point divided by the current at the same point. Thus if the standing waves of current and voltage indicate that, at a point, the voltage is high and current low the line at that point has a high impedance. The reactance curve shows the reactance relationships along the sections and indicates whether the reactance at any point is purely resistive. A comparison of the reactance curves with the impedance curves also shows the points where the impedance of the line is purely resistive. This is due to the reactances having cancelled themselves out or to their becoming zero. The circuit symbols indicate the conventional circuit which is equivalent to a resonant line of that particular length.

The quarter-wavelength section exhibits the characteristics of a conventional series-resonant circuit with high

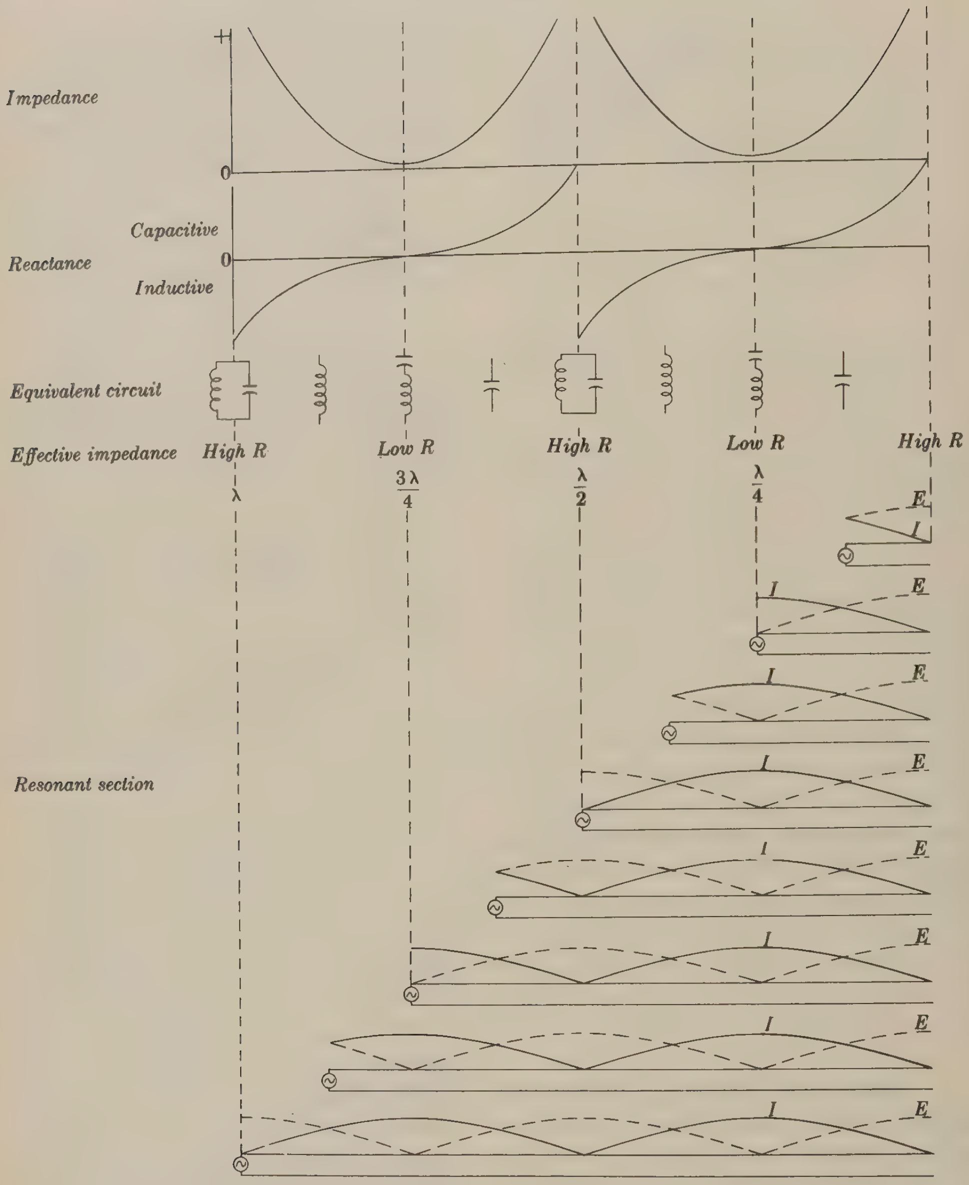


FIG. 10-30 Relationship between voltage, current and impedance for various lengths of an open-ended resonant line.

current, low voltage and an impedance minimum. Thus the generator looking into a quarter-wave open-ended section sees a very low impedance or in effect the generator sees a short circuit. Similarly the $\frac{3}{4}\lambda$ section presents the same load to the generator. In general, an open-ended resonant-line section which is an odd number of quarter wavelengths long exhibits the characteristics of a conventional series-resonant circuit and so looks like a short circuit to the generator.

The open-ended resonant section may be used also to act as nearly pure capacitance or inductance. Fig. 10-30 shows that an open-ended line section less than a quarter wavelength long acts as a

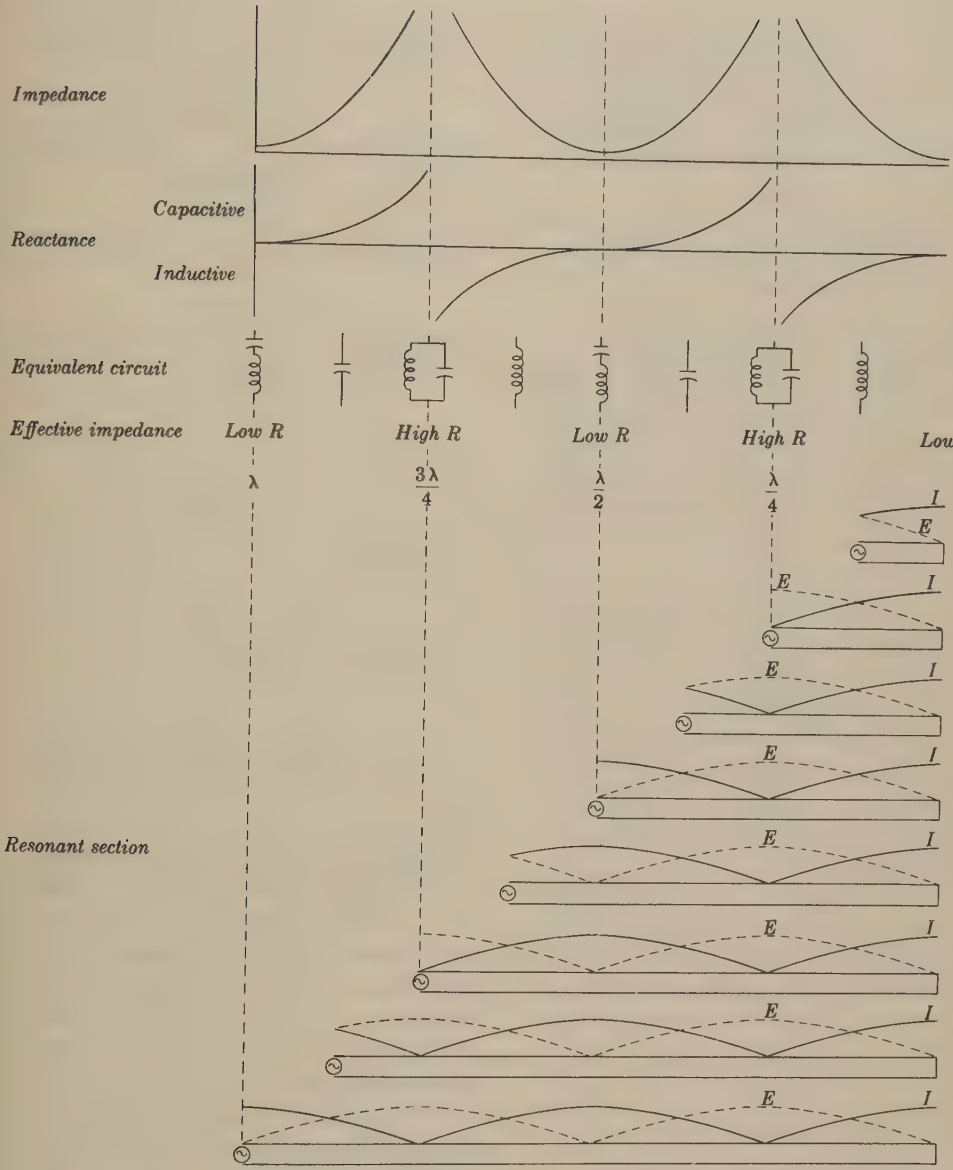


FIG. 10-31 Relationship between voltage, current and impedance for various lengths of a short-circuited resonant line.

capacitance; also a section from $\frac{\lambda}{4}$ to $\frac{\lambda}{2}$ long acts as an inductance and so on. In addition a $\frac{\lambda}{8}$ open-ended section acts as a capacitive reactance numerically equal to the characteristic impedance Z_o and a $\frac{3}{8}\lambda$ section acts as an inductive reactance numerically equal to Z_o.

B. Short-Circuited Resonant-Line Section

Fig. 10-31 illustrates the relationship between voltage, current and impedance for various sections of a short-circuited resonant line.

The quarter-wavelength section exhibits the characteristics of a conventional parallel-resonant circuit with low current, high voltage and an impedance maximum. Thus the generator looking into a quarter-wave short-circuited section sees a very high impedance or, in effect, an open circuit. Similarly the $\frac{3}{4}\lambda$ section presents the same load to the generator. In general a short-circuited resonant-line section which is an odd number of quarter wavelengths long exhibits the characteristics of a parallel-resonant circuit and so looks like an open circuit to the generator.

The half-wavelength section exhibits the characteristics of a conventional series-resonant circuit with high current, low voltage and a minimum impedance. Thus the generator, looking at a half-wave shorted section, sees a very low impedance or, in effect, the generator sees a short circuit. Similarly the full wavelength section presents the same load to the generator. In general, a short-circuited resonant-line section which is an even number of quarter wavelengths long exhibits the characteristics of a conventional series-resonant circuit and so looks like a short circuit to the generator.

In addition to acting as conventional resonant circuits, resonant short-circuited lines may be also used to act as nearly pure capacitance and inductance. For example, a short-circuited line less than $\frac{\lambda}{4}$ acts as an inductance, from $\frac{\lambda}{4}$ to $\frac{\lambda}{2}$ long it acts as a capacitance, and so on, as indicated in the diagram. A $\frac{\lambda}{8}$ shorted-line section acts as an inductive reactance equal to the characteristic impedance Z_o.

10.11 Characteristics of quarter-wavelength and half-wavelength line sections.

A. The Quarter-Wave Section

It has been seen from Fig. 10-30 and 10-31 that in the case of the quarter-wave section, different terminations place different loads on the generator. The impedance presented to a generator by a quarter-wave section with different terminations is shown in Fig. 10-32. In each case it is important to note that the quarter-wave section always inverts its load, that is, a short circuit appears as an open circuit to the generator; a capacitance appears as an inductance; a low resistance looks like a high resistance. Also, any line which is an odd number of quarter wavelengths long acts like a quarter-wave section.

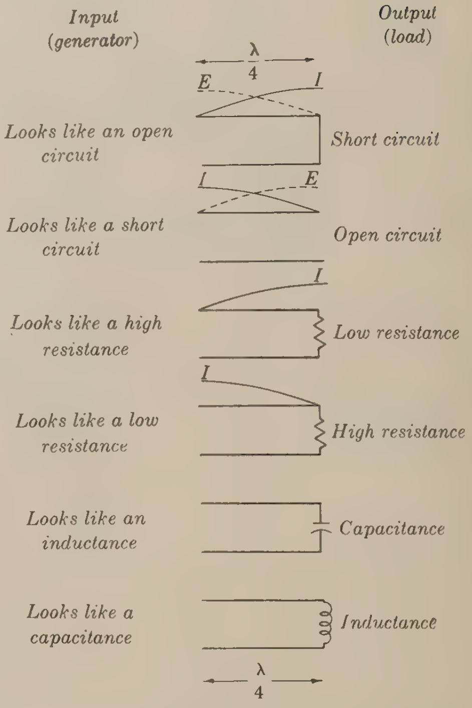


FIG. 10-32 Impedance presented to a generator by a quarter-wave section with different terminations.

B. The Half-Wave Section

From Fig. 10-30 and Fig. 10-31 it has been seen that an open-ended half-wave section presents a high impedance to the generator and a short-circuited half-wave section presents a low impedance to the generator. In effect the important fact is that a half-wave section always presents to its generator a load equal

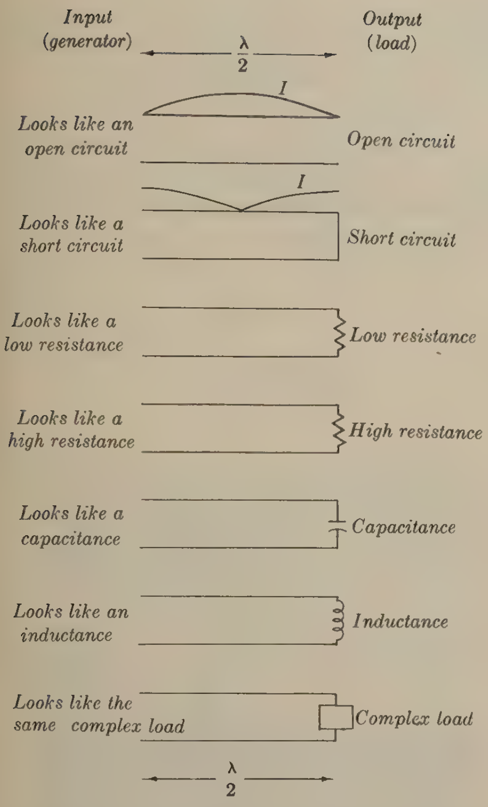


FIG. 10-33 Impedance presented to a generator by a half-wave section with different terminations.

to its termination. In other words, a half-wave section repeats its load. The impedance presented to a generator by half-wave sections with different terminations is shown in Fig. 10-33. It follows also that a line any multiple of a half wavelength long also repeats its load.

10.12 Construction of transmission lines and determination of characteristic impedance. There are several distinct types of transmission lines in common use. These fall readily into the following six distinct classifications:

(i) Single-wire line
(ii) Two-wire line
(iii) Twisted pair
(iv) Shielded pair
(v) Multi-conductor
(vi) Concentric (coaxial) cable.

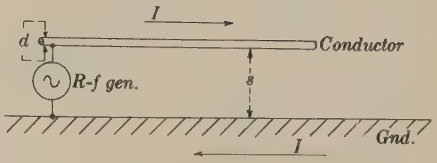


FIG. 10-34 Single-wire transmission line.

A. The Single-Wire Line

The simplest type of transmission line is the single-wire line (Fig. 10-34). The single-wire line is not used in Service applications to any great extent. This is due to the fact that losses by radiation are great because the use of a ground return does not provide out-of-phase radiation to cancel the radiation from the single wire. Further, the conductivity of the ground plays an important part, since rocky, dry or sandy soil causes severe attenuation. In addition it is difficult to prevent harmonics entering such a system and being radiated with resulting undesirable interference. In emergencies a single-wire line could be employed as it will operate over a fair range of frequencies without serious losses.

The characteristic impedance of a single-wire line with air dielectric is calculated from the formula:

$$Z_o = 138 \log_{10} \frac{2s}{d}$$

where Z_o = characteristic impedance of line in ohms
s = distance separating centre of conductor from ground in any linear unit
d = diameter of wire conductor, in the same linear unit.

B. The Two-Wire Line

The two-wire parallel conductor transmission line is the basic type. The two wires of the line are maintained parallel at a fixed distance by means of insulating spacers or spreaders at suitable intervals,

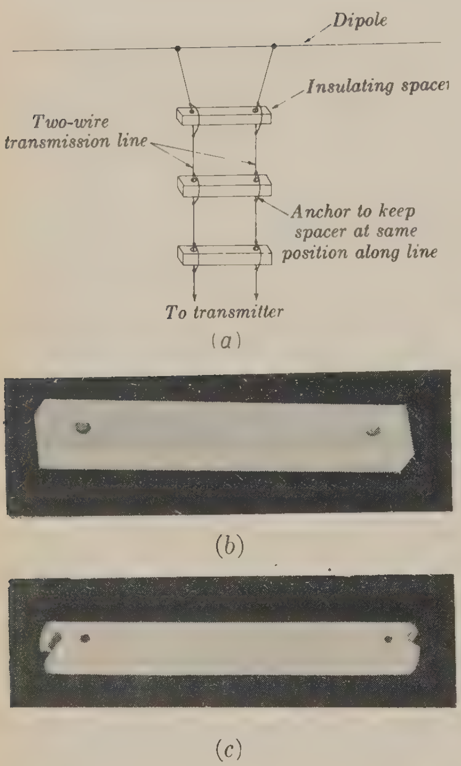

Dipole
Insulating spacer
Two-wire transmission line
Anchor to keep spacer at same position along line
To transmitter
(a)
(b)
(c)

FIG.10-35 Two-wire transmission line (a) Construction of line (b) Typical spacer insulators.

as in Fig. 10-35 (a). Typical insulators which are used as spacers are shown in Fig. 10-35 (b) and (c). The two-wire line is economical, easy to construct and efficient. It is used for commercial power lines, telephone and telegraph lines and to connect transmitter to aerial and receiving aerial to its receiver. Two-wire lines are spaced from 2 to 6 inches apart for frequencies of 2814 Mc/s or lower. Four inches is maximum spacing for frequencies above 2814 Mc/s.

A typical transmission line used to transfer r-f energy from an l-f communications transmitter over a considerable distance to an aerial is shown in Fig. 10-36 (a). The method of effecting a turn in the line is shown in Fig. 10-36 (b).

The two-wire line is not used in shipboard applications to any great extent due to the difficulty of maintaining adequate insulation between the wires. This is due to the fact that an open-air line is easily accessible to salt spray.

The characteristic impedance of a transmission line can be shown to be approximately equal to the square root of the ratio of inductance to capacitance of the line, that is,

$$Z_o = \sqrt{\frac{L}{C}}$$

where Z_o = characteristic impedance of line in ohms

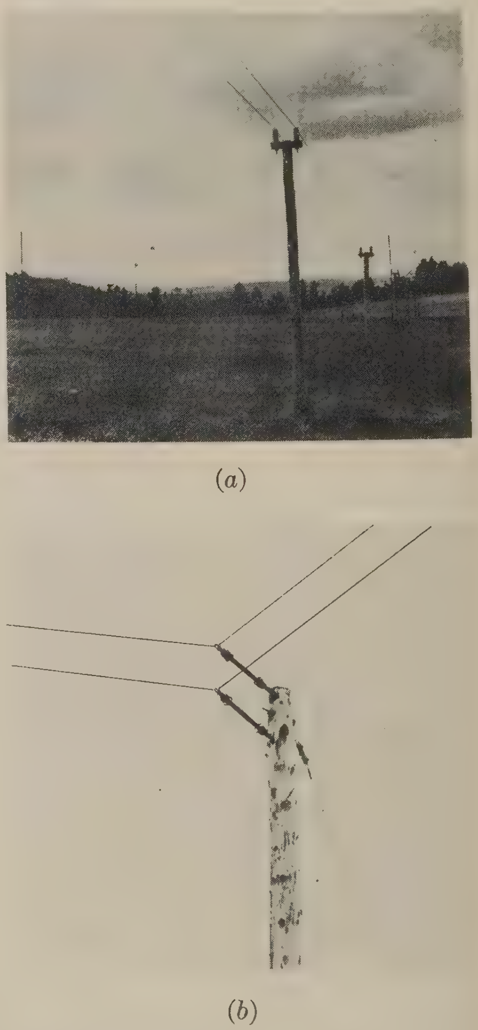


FIG.10-36 Two-wire r-f transmission line (a) A straight run (b) A turn.

and L = inductance of line per unit length in henries
C = capacitance of line per unit length in farads.

For a two-wire transmission line it can also be shown mathematically that the value of inductance and capacitance per unit length of this line can be calculated from the formulas,

$$L = k_1 \log_{10} \frac{2s}{d}$$

$$C = k_2 \frac{1}{\log_{10} \frac{2s}{d}}$$

where s = distance separating the centres of the conductors
d = the outer diameter of one conductor
and k_1 and k_2 are constants.

It can be seen from the formulas that an increase in the separations of the conductors increases the inductance, decreases the capacitance, and hence increases the value of the ratio of inductance to capacitance $\left(\frac{L}{C}\right)$ and so the characteristic impedance. This is partly due to the fact that two wires, closely spaced, carrying currents in opposite directions, have a very low effective inductance and partly because the effective capacitance is relatively high if the plates of the capacitor, that is the two wires, are close together. Similarly if the wire diameter d is made smaller, in effect the size of the capacitor plates are made smaller and the capacitance is also made smaller. There is also an increase in inductance but the charge affects the capacitance more than the inductance and hence the ratio of $\frac{L}{C}$ is effectively increased and so the characteristic impedance is increased.

In addition a change in dielectric material may also increase the capacitance and so decrease the characteristic impedance.

The characteristic impedance of a two-wire line with air dielectric is given by the formula,

$$Z_o = 276 \log_{10} \frac{2s}{d}$$

where s = distance separating centres of conductors in any convenient linear unit
d = diameter of one conductor in same linear unit.

Example 1. Find the characteristic impedance of a two-wire line made of copper tubing $\frac{1}{4}$ inch in diameter spaced 3 inches apart.

$$\begin{aligned} Z_o &= 276 \log_{10} \frac{2s}{d} \\ &= 276 \log_{10} \frac{6}{\frac{1}{4}} \\ &= 276 \log_{10} 24 \\ &= 276 \times 1.38021 \text{ (from log tables)} \\ &= 380.94 \end{aligned}$$

Therefore, the characteristic impedance of the line is 381 ohms.

Fig. 10-37 shows a characteristic impedance chart for two-wire air-dielectric lines. This chart gives the characteristic impedance for lines consisting of various types of wire and spacing of the conductors. Thus the characteristic impedance for a given line can be read off this chart directly. Also the required physical dimensions of a line required to provide a desired characteristic impedance can be read off directly.

Example 2. From the characteristic impedance chart (Fig. 10-37) find Z_o of a two-wire line of copper tubing $\frac{1}{4}$ inch in diameter spaced 3 inches apart.

On the horizontal axis locate the vertical line corresponding to 3-inch spacing. Move vertically along this line to the point of intersection with the curve for transmission lines made of $\frac{1}{4}$-inch tubing. Opposite to this point read off the characteristic impedance in the vertical axis.

Thus, Z_o = 380 ohms.

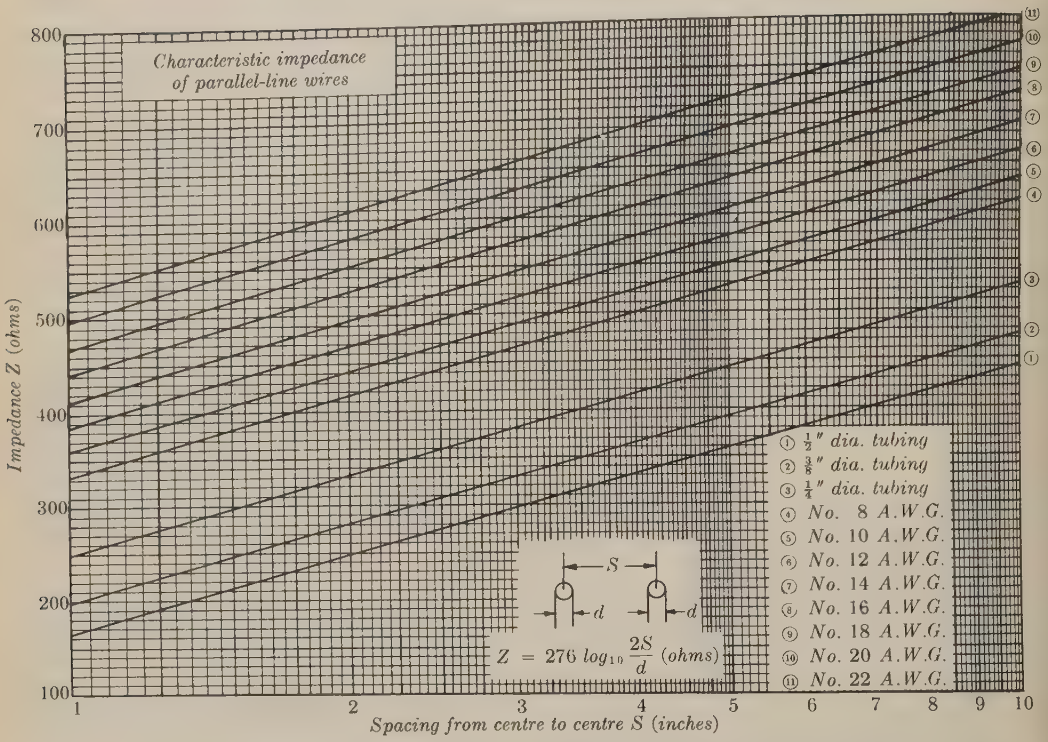


FIG. 10-37 Characteristic impedance chart for two-wire air-dielectric line.

Example 3. From the characteristic impedance chart (Fig. 10-37) find the spacing required in construction of a two-wire line of No. 18 wire to provide a characteristic impedance of 600 ohms.

Locate the intersection of the 600-ohm line and the curve representing No. 18 wire.

Opposite this point read off the required spacing from the horizontal axis.

Required spacing = 3 inches.

C. The Twisted Pair

The *twisted-pair* transmission line usually consists of two rubber- and cloth-covered wires twisted together as shown in Fig. 10-38.

FIG. 10-38 A twisted-pair transmission line.

It has been seen (Sec. 10-8) that if the separation between the wires of a two-wire line is kept small with respect to the wavelength there is very little radiation from the lines. For a twisted pair the separation between the wires is very small and the radiation is small. The resultant radiation from the two-wire line depends upon the phase difference set up as a result of the separation. Because of the frequent twists in a twisted pair the effect of this phase difference due to wire separation is further reduced. Hence the radiation for a twisted-pair transmission line is reduced and the pick-up of a twisted-pair line used to feed a receiver is also reduced. This effect can be understood by a consideration of Fig. 10-39. An electromagnetic wave cuts the twisted-pair and currents are induced in each wire in the same direction. The current which results from currents induced in sections *A* and *B* depends on the phase difference

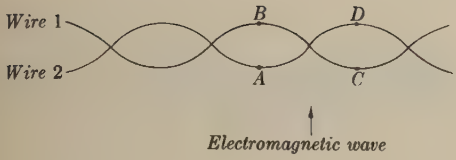


Fig.10-39 Diagram to aid in showing why radiation from a twisted-pair line is less than from a spaced two-wire line.

introduced by the distance between A and B. Thus the current in wire 1 leads that in 2 by some value. However, since the wires are now twisted the currents induced in section C and D are out of phase but the current in wire 2 leads that in wire 1 by a phase angle which is approximately equal to the previous phase difference. The same is true for all other sections and the resulting overall phase difference is reduced to a small value and the pick-up current is at a small value. For the same reason the radiation from a twisted pair is reduced to a very small value.

In transmission applications the twisted pair is used as a non-resonant line. A two-wire line composed of twisted rubber-covered wire can be constructed to have an impedance approximately equal to that at the centre of a half-wave aerial, that is 73 ohms. Thus this type of line (Fig. 10-40) can be connected directly to the centre of the aerial. Any matching discrepancy which may exist between the line and aerial impedance can be compensated for by a slight fanning out of the line where it connects to the two halves of the aerial. The twisted line is a convenient type to use, since it is easy to install and the r-f voltage on it is low because of its low impedance. This makes insulating a relatively easy matter. The twisted pair is used as a non-resonant line for low-frequency transmission. It is not used at the higher frequencies because of the high losses occurring in the rubber insulation. These excessively high losses also make its use practical over only very short distances at low frequencies. Its chief advantage is that it may be used where more efficient lines are not possible because of space and mechanical considerations.

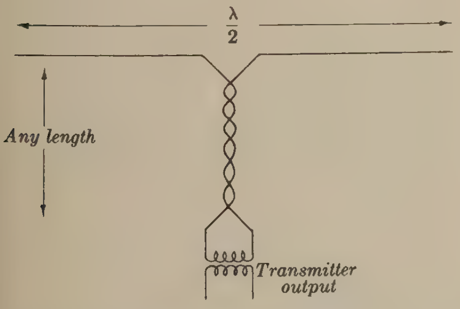


Fig.10-40 Method of connecting a twisted-pair line to a dipole.

The twisted pair is particularly useful to connect aerial to receiver as no great dielectric strength is required of the insulation in such applications. Further, full advantage is taken of the low pick-up of this line.

The characteristic impedance of a twisted-pair line may be determined if a knowledge of the effect of the dielectric is known.

D. The Shielded Pair

The *shielded pair* consists of two parallel conductors separated and maintained a fixed distance apart by means of insulating spacers such as polystyrene or surrounded entirely by an insulating dielectric material. A copper-braided tubing surrounds the insulating material to form a shield for the wires. This is in turn surrounded by a rubber coating to protect the line against moisture and friction.

A very simple shielded pair is shown in Fig. 10-41 employing two conductors and a shield with cotton covering forming the insulation. The cable shown in Fig. 10-42 is a high-frequency low-loss cable and is constructed of two No. 18 solid

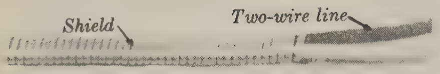


FIG. 10-41 A shielded pair of simple construction.

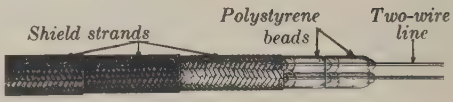


FIG. 10-42 An h-f low-loss shielded pair.

tinned copper wires strung with polystyrene beads. The ball and socket design of the beads permits bending the cable without breaking the polystyrene and without baring the conductors; it also keeps the conductors parallel even around bends. The cable shield is usually made of several strands of copper-wire mesh. The characteristic impedance of such cable is of the order of 150Ω.

The advantage of the twisted pair is that the two wires are balanced to ground, that is, the capacitance between each wire and ground is uniform along its entire length. This balance is obtained by means of the grounded shield. This shield thus further reduces radiation and pick-up of stray fields. This can be explained by a consideration of the shielding action. If an unshielded line is not to radiate the currents in each conductor must be equal in amplitude so that equal and opposite fields are set up which cancel. This requirement is obtained only if the line is situated in the clear, away from all grounded metal objects, and if the distance between the wires is small. However, if the line runs near some grounded conducting surface one of the wires is nearer to it than the other. As a result a capacitance exists between that wire and the grounded object which differs from the capacitance between the other wire and the object producing an unbalance. These capacitances act as parallel conducting paths for each wire of the line and the current flow through one is greater than through the other. In this way the current in one wire of the line is made different from that in the other. This inequality results in incomplete cancellation of radiation. The shield of the shielded-pair reduces the effect of surrounding metallic objects by maintaining a fixed balanced capacitance to ground for each wire. In this way the shield reduces radiation losses and pick-up to a minimum.

E. Multi-Conductor Lines

1. THE FOUR-WIRE LINE.

Another method of obtaining balance in a line without the use of a shield is by means of four wires. In the four-wire line (Fig. 10-43) the opposite wires are connected to each other. In this way any metallic object near one wire is also close to one wire of the other pair. Thus the tendency towards unbalance is reduced and the current in one pair is nearly the same as in the other. As a result the radiation is less than for a two-wire line but greater than for a shielded pair.

The characteristic impedance of the four-wire line with air dielectric is given by,

$$Z_o = 138 \log_{10} \frac{2s}{d} - 21.$$

where s = spacing between adjacent wires

d = diameter of wire.

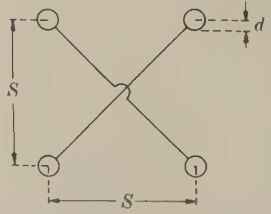


FIG. 10-43 Cross-section of a four-wire line.

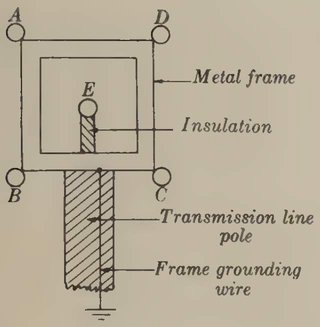


FIG.10-44 Cross-section of a five-wire line.

2. The Five-Wire Line.

Some shore installations make use of a five-wire line to reduce unbalance and provide an even better shielding effect than that provided by the four-wire line. The five-wire line (Fig. 10-44) consists of four outside wires, *A*, *B*, *C* and *D* and a central conductor *E*. The four outside wires are supported by means of a square metal frame mounted on the transmission line poles. This frame is also grounded by a length of wire from frame to ground, thus placing the outer conductors at ground potential to improve the line balance. The inner conductor *E* is supported on insulation inside the frame at each pole.

F. The Concentric or Coaxial Line

1. Rigid Coaxial Line.

The development of the five-wire line led to the construction of the concentric or coaxial line which provides very efficient operation even at extremely high frequencies. The coaxial line (Fig. 10-45) consists of a cylindrical conductor inside of, and coaxial with, an outer tubular conductor. The inner conductor of a rigid coaxial line is held at the centre of the outer conductor by insulating spacers, Fig. 10-46 (a), situated at regular intervals. The inner conductor may be solid, tubular or multi-strand. The spacers are made of pyrex, polystyrene, steatite or similar material with good mechanical strength, good insulating properties and low power loss characteristics.

The main purpose of the coaxial line is to reduce radiation losses to a minimum and at the same time reduce other line

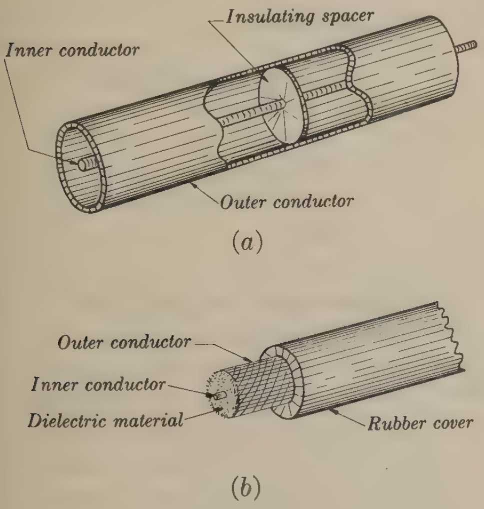


FIG.10-45 Construction of coaxial line (a) Rigid line with insulating spacers (b) Flexible line with solid dielectric.

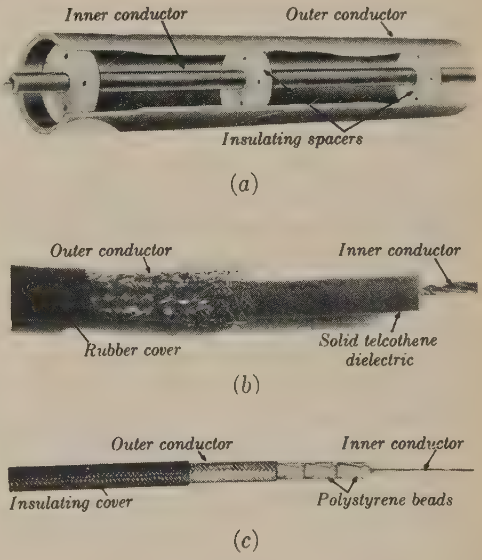


FIG.10-46 Coaxial transmission lines (a) Rigid line (b) Flexible line using beaded insulators (c) Flexible solid-dielectric line.

losses to an extremely low value. The latter is effected by means of the shape and size of the conducting surfaces. The former is effected by the complete shielding of the inner conductor. The outer conductor is operated at zero potential by grounding it conveniently usually at its external supports. This conductor then provides a complete shield for the inner conductor and radiation losses are extremely small.

The concentric line finds wide application in radar equipments and it is used for frequencies ranging from 100 to 3000 Mc/s. Bolted or screwed couplings are used to join sections of a rigid coaxial line.

One of the main shipboard disadvantages of concentric line is that it must be kept dry to prevent excessive leakage between the conductors. In practice the lines are sometimes filled with dry nitrogen maintained at a pressure of 8 to 30 lbs. to prevent condensation and keep the line dry. The pressure ensures that all leakage is outward. Thus all such lines and joints must be gas-tight. This provides one of the greatest difficulties since it is difficult to maintain tight joints due to vibration. Vibration is also responsible for many cracked seams in coaxial feed lines.

In general, the main advantage of the coaxial line is the excellent shielding that is obtained and the line's increased efficiency. However in shipboard installations its greatest advantage is its immunity to weather conditions.

The characteristic impedance of a rigid air dielectric concentric line is given by the formula,

$$Z_o = 138 \log \frac{D}{d}$$

where Z_o = characteristic impedance in ohms

D = inner diameter of outer conductor in any convenient linear unit

and d = outer diameter of inner conductor in the same unit.

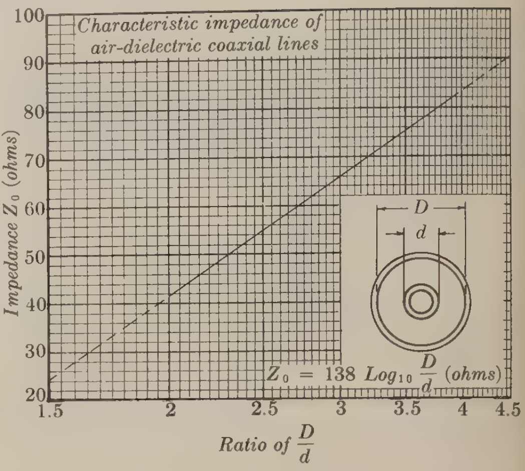


FIG. 10-47 Characteristic impedance chart for rigid coaxial lines of various dimensions.

Fig. 10-47 is a chart which gives the characteristic impedance of a concentric line of given physical dimensions. The chart also provides a method of selecting line dimensions for a desired characteristic impedance.

Example. Find the characteristic impedance of a line having $D = 0.51$ inches, $d = 0.16$ inches.

USING CHART

$$\text{Ratio of } \frac{D}{d} = 3.2$$

Thus from the chart

$Z_o = 70$ ohms.

USING FORMULA

$$\begin{aligned} Z_o &= 138 \log 3.2 \\ &= 138 \times 0.50515 \\ &= 69.7 \end{aligned}$$

Z_o = approximately 70 ohms.

2. THE FLEXIBLE COAXIAL LINE.

The solid dielectric coaxial transmission line is shown in Fig. 10-45 (b) and Fig. 10-46 (b). The dielectric is stabilized polyethylene or telcothyne material or a mixture of polyisobutylene and other materials. The dielectric is applied tightly around the centre conductor. The outer conductor or shield consists of a plain or tinned copper mesh and it is applied to fit tightly around the insulating dielectric. If the line is to be used under conditions where shielding is critical, a double-wire braid or a lead sheath is used. The cable is enclosed in a tough, flexible protective covering over the outer conductor or shield.

The solid dielectric lines are much more flexible than gas-filled lines and so can be used for many purposes where it is impractical to use other types. The solid dielectric line is much easier to connect and after connection it provides a nearly continuously uniform line throughout its length.

With a dielectric other than air the formula for the characteristic impedance must take into account the dielectric constant of the material, since the capacitance of the line is changed with a change in dielectric. Thus the formula becomes,

$$Z_o = \frac{138 \log \frac{D}{d}}{k}$$

where k = constant which depends on factors involving the dielectric.

The constant k is greater than unity and thus it is apparent from this formula that the characteristic impedance of a solid dielectric line is lower than that of a corresponding air dielectric line.

A second common type of flexible coaxial line is shown in Fig. 10-46 (c). This line employs beaded spacers to hold the inner conductor in position. Beaded insulation is preferable to a solid dielectric because dielectric losses are less. This is due to the fact that with beaded insulation there is less solid dielectric present. The beads are often made of ceramic materials but at frequencies of the order of 3000 Mc/s they are made of polystyrene. Polystyrene beads break down from high voltages. This breakdown occurs at the beads as a result of the concentration of the electric field in the minute air spaces between the beads causing the air to ionize first. Carbonization of the polystyrene by the ionization is a permanent fault which is not self healing after the high voltage is removed. A further disadvantage in some installations is that dust and condensed moisture on the beads provide a leakage path.

10.13 Applications of resonant sections.

A. Transmission-Line Oscillators

An important application of resonant transmission-line sections is their use as circuit components for high-frequency oscillators. Since a short-circuited quarter-wave section exhibits the characteristics of a conventional parallel resonant circuit, such a section may be used to replace the ordinary coil-capacitor tank circuits of conventional oscillators. A typical push-pull transmission-line oscillator designed to operate at 120 Mc/s is shown in Fig. 10-48 and Fig. 10-49. The anode tank circuit consists of quarter-wave brass tubes. This resonant-line section provides required lumped inductance and capacitance together with the additional capacitance of C_1 to make the line resonant at 120 Mc/s. The large conducting surface of the tank lines provide relatively low high-frequency ohmic resistance. Further, the rigidity of construction minimizes the effect of mechanical vibration on the frequency. In addition, the currents in the two parallel rods are at any instant in opposite directions and hence their magnetic fields cancel. Thus this reduces radiation losses and lowers radiation resistance. Since the resistance losses are very low the Q of the line, $\frac{\omega L}{R}$, is high.

The Q of a two-wire line is a maximum if

$$\frac{s}{d} = 6.19$$

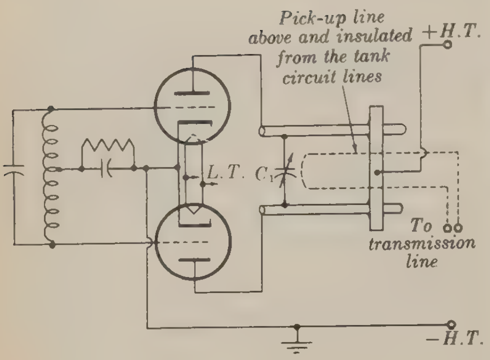

Pick-up line above and insulated from the tank circuit lines
+H.T.
L.T.
C_1
To transmission line
−H.T.

FIG. 10-48 Circuit of a push-pull transmission-line oscillator designed to operate at 120 Mc/s.

where s = the distance separating centres of conductors
d = the outer radius of one conductor.

The Q of a concentric line is a maximum if

$$\frac{D}{d} = 4.22$$

where D = the inner diameter of the outer conductor
d = the outer diameter of the inner conductor.

The advantages of resonant-line sections used as tuned circuits in oscillators are high Q, low losses and efficient oscillation at frequencies above those which are readily attainable with conventional tank circuits.

Transmission-line sections which are used as tuned-circuit elements are often referred to as Lecher lines.

B. The Quarter-Wave Shorted Section as an Insulator

The fact that a quarter-wave section, with its output end short-circuited looks like an infinite impedance to the source makes such a section very useful as an insulator. For example, when a two-wire non-resonant line is to be run from a transmitter to an aerial as shown in Fig. 10-50 it is often convenient to use quarter-wave sections as insulators in effecting a turn in the transmission line.

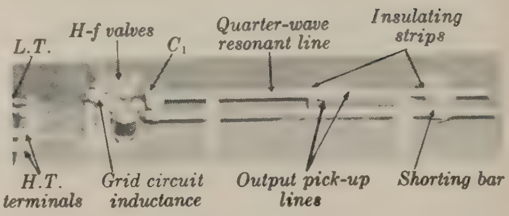


FIG. 10-49 Photograph of the oscillator of Fig. 10-48.

The quarter-wave section connected to the line at AA' is short-circuited at BB' Thus the energy being transferred along the transmission line to the aerial sees an infinite impedance presented by the quarter-wave section and therefore no current is diverted from the main line. Thus the non-resonant transmission line is insulated from the deck.

A quarter-wave section may also be used as an insulator to support and space a parasitic reflector one quarter wavelength from a driven dipole (Fig. 10-51). The purpose of the reflector is to cause all the energy from the dipole to be radiated away from the dipole on the side opposite to the reflector. When a concentric line is used as a feed line a quarter-wave section of the line may be employed as shown.

Another application of the quarter-wave shorted section as an insulator is shown in Fig. 10-52. It has been found

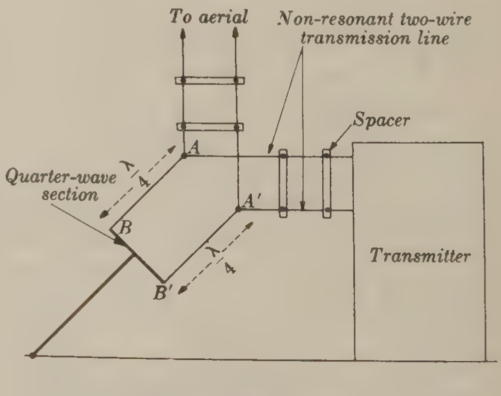


FIG. 10-50 Use of a quarter-wave section as an insulator in effecting a turn in a transmission line.

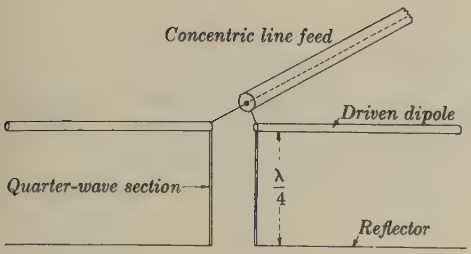


FIG. 10-51 Use of a quarter-wave section to support and space a parasitic reflector.

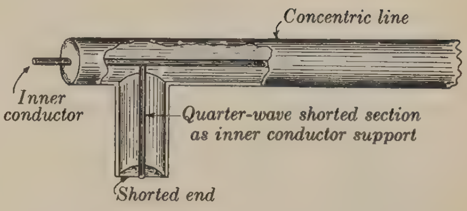


FIG. 10-52 Use of a quarter-wave section to support and insulate the inner conductor of a coaxial line.

that the ceramic disks used as spacers to hold the centre conductor of a concentric line in position are not efficient insulators at very high frequencies around 3000 Mc/s. In some applications therefore, where the line is operated at one frequency only quarter-wave shorted sections are placed along the line as shown in the diagram to hold the inner conductor in position.

C. Quarter-Wave Sections as Wave Filters

Quarter-wave line sections can be used as wave filters to suppress even harmonic frequencies of the fundamental frequency. Suppose a transmitter which is operating at 100 Mc/s on test is found to radiate harmonics of 200 Mc/s and 300 Mc/s in excessive amount. A quarter-wave section can be used as a wave filter to reduce this harmonic radiation.

From Fig. 10-31 it can be seen that a quarter-wave shorted section acts as a parallel-resonant circuit and so offers maximum impedance to its fundamental frequency, for example, 100 Mc/s. At the frequency of the 2nd harmonic, twice the fundamental, the quarter-wave section now becomes a half-wave section and from Fig. 10-31 it can also be seen that a shorted half-wave section offers zero impedance at its input end. Similarly at the frequency of the 4th harmonic the quarter-wave section becomes a full-wave section and offers zero impedance to this frequency. This principle is made use of in Fig. 10-53 to eliminate even harmonics. The resonant wave filter shown is one quarter wavelength long at the fundamental frequency and offers a maximum impedance to current at this frequency. However, at the 2nd harmonic frequency it is one half wavelength long and the section offers almost zero impedance to the current at this frequency and effectively grounds it. Similarly the 4th and other even harmonics are effectively grounded and do not reach the aerial to be radiated. The quarter-wave filter may be connected anywhere across the non-resonant transmission line.

This method of wave trapping cannot be used for the odd-multiple harmonics. The quarter-wave section at the fundamental frequency now becomes a $\frac{3}{4}\lambda$ section at the frequency of the 3rd harmonic. Thus this section also presents a high impedance at this and all

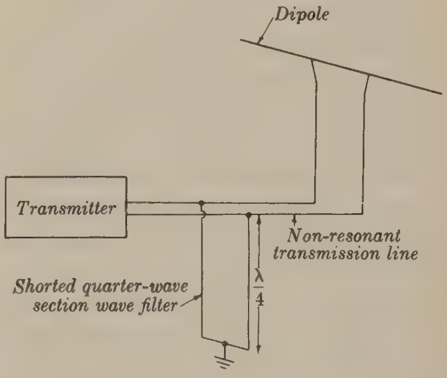

Dipole
Transmitter
Non-resonant transmission line
Shorted quarter-wave section wave filter
λ/4

Fig. 10-53 Use of a shorted quarter-wave section as a wave filter.

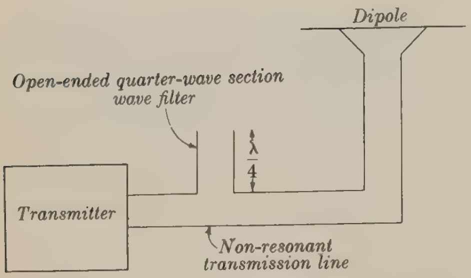

Dipole
Open-ended quarter-wave section wave filter
$\frac{\lambda}{4}$
Transmitter
Non-resonant transmission line

Fig. 10-54 Use of an open-ended quarter-wave section as a wave filter.

other odd harmonics and no filtering action results.

An open-ended quarter-wave section may also be used as a wave filter. Such a section is inserted in the line as shown in Fig. 10-54. An open-ended quarter-wave section presents a load that looks like a short circuit to the fundamental frequency, However, the 2nd harmonic sees an open circuit or infinite impedance since at this frequency the section is one half wavelength long. For a similar reason, an infinite impedance is presented to each even harmonic frequency. Thus this quarter-wave open-ended filter passes the fundamental frequency and each odd harmonic along the line.

D. Shorted Sections as Impedance-Matching Devices

1. Resonant Line to Non-resonant Line.

A quarter-wave shorted section may be used to match a resonant line to a non-resonant line as shown in Fig. 10-55 (a). To make the input line non-resonant it must be terminated in its characteristic impedance and this is accomplished by correct setting of the tapping points, 3, 4. This can be understood by a consideration of the equivalent conventional resonant circuit, Fig. 10-55 (b). If the line voltage is applied across 3, 4 the circuit remains, for all practical purposes, in resonance but the impedance presented to the line is lower. Hence by correct selection of tapping points 3, 4 it is possible to terminate the line in a purely resistive load matched to its characteristic impedance.

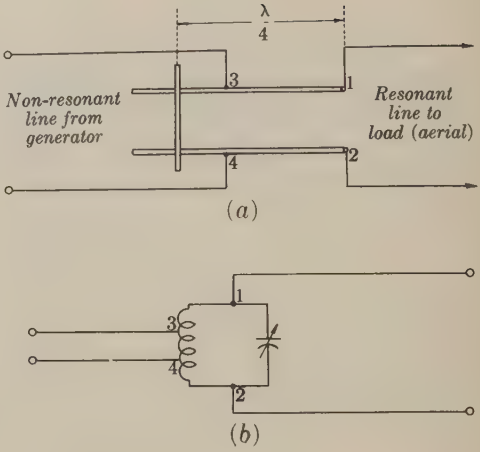

$\frac{\lambda}{4}$
3
1
Non-resonant line from generator
Resonant line to load (aerial)
4
2
(a)
1
3
4
2
(b)

Fig. 10-55 Use of a shorted quarter-wave section to match a non-resonant line and a resonant line (a) Method of connection (b) Equivalent circuit.

Thus, in Fig. 10-55 (a), in order to terminate the non-resonant line in its characteristic impedance, it is only necessary to adjust the position of tapping points 3, 4. If the tapping points are set at the shorting bar the terminating impedance is zero. As the tapping points are moved along the quarter-wave section the impedance increases. The impedance presented to the non-resonant line is resistive as the section is tuned. The section is tuned before the tapping points are adjusted. The output to the resonant line is taken from points 1 and 2 where the impedance and voltage are a maximum.

2. Receiver to Input Line.

The method of matching a receiver to an r-f line depends upon the type of input circuit used in the receiver. One common input circuit employs a quarter-wave shorted transformer to match the coaxial input line to the grid circuit as shown in Fig. 10-56 (a). The simplified equivalent circuit is shown in Fig. 10-56 (b).

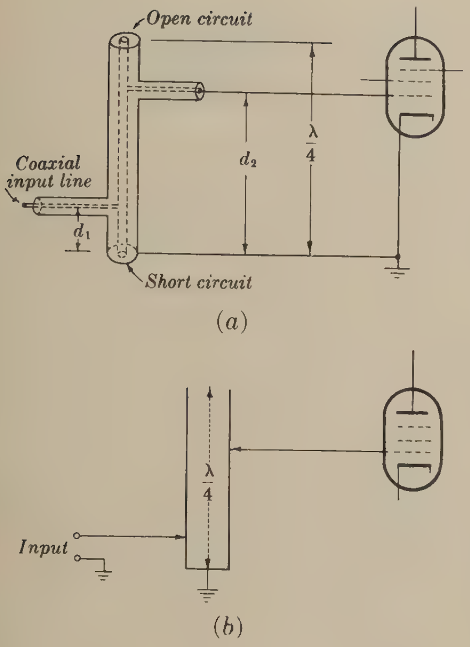


FIG. 10-56 Use of a shorted quarter-wave transformer to match a coaxial input line to a valve grid circuit (a) Method of connection (b) Simplified circuit.

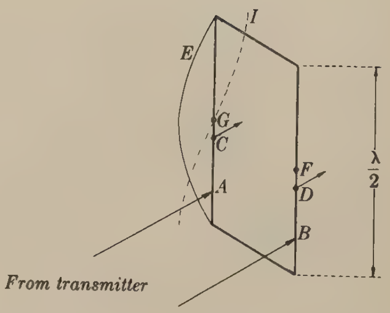


FIG. 10-57 Use of a half-wave section shorted at both ends to match an aerial to a non-resonant line.

For maximum signal input the distances d_1 and d_2 are chosen so that the coaxial line is terminated in its characteristic impedance. In practice the coaxial line is terminated in a resistance which is slightly greater than the characteristic impedance of the line to improve the signal-to-noise ratio. Thus the characteristic impedance of the line which is relatively low is transformed to a high impedance to match the high input impedance of the valve.

3. HALF-WAVE SECTION SHORTED AT BOTH ENDS.

A half-wave section shorted at both ends used to match an aerial to a non-resonant line is shown in Fig. 10-57. The standing waves of current and voltage are also shown in the diagram. The line from the transmitter sees the impedance Z_{AB} which is equal to the value of $\frac{E}{I}$ at that point. Thus the tapping points, A and B, may be adjusted so that the line is terminated in its characteristic impedance. The greatest impedance occurs at GF, the centre points of the half-wave section, and the minimum impedance occurs at the shorting bars. Since the upper half repeats the impedance of the lower half there are two positions on the half-wave section which have the same impedance. Either of these positions may be used, the only difference being that the current in one half is 180° out of phase with that in the other half. The aerial is connected across C and D and the position of the tapping points adjusted to match the aerial impedance for maximum transfer of energy.

E. Open-Ended Sections as Impedance-Matching Devices

1. THE QUARTER-WAVE TRANSFORMER.

An open-ended quarter-wave section, usually referred to as a quarter-wave transformer, can be used to match one transmission line to another transmission line which has a different characteristic impedance, or to match a transmission line of one characteristic impedance to an aerial with a different impedance.

It can be shown mathematically that the input impedance, Z_s, of an open-ended quarter-wave section (Fig. 10-58) is equal to the square of the characteristic

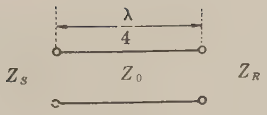


FIG. 10-58 A quarter-wave transformer.

impedance divided by the impedance Z_R at its receiving end.

Thus,
$$Z_S = \frac{Z_o^2}{Z_R}$$

where Z_S = the input impedance

Z_o = the characteristic impedance

Z_R = the terminating impedance or load.

From this formula it can be readily seen that the quarter-wave transformer can be put to practical use when it is necessary to match a line of one impedance to a load of a different impedance (Fig. 10-59). To do this the characteristic impedance of the quarter-wave transformer which forms the matching section can be calculated from the formula,

$$Z_o = \sqrt{Z_S \times Z_R}$$

where Z_o = the characteristic impedance of the quarter-wave matching section

and Z_S = the input impedance, that is, the characteristic impedance of the transmission line

Z_R = the terminating impedance, that is, the impedance of the dipole.

Example 1. Find the characteristic impedance of a quarter-wave transformer required to match a 500Ω transmission line to a 72Ω dipole (Fig. 10-59).

$$\begin{aligned} Z_o &= \sqrt{Z_S \times Z_R} \\ &= \sqrt{500 \times 72} \\ &\doteqdot 190 \end{aligned}$$

Therefore, the characteristic impedance of the matching section is 190 ohms. Thus the transmission line sees an impedance which is equal to $\frac{Z_o^2}{Z_R}$, that is, $\frac{190^2}{72}$ or 500 ohms.

Also the dipole sees an impedance which is equal to $\frac{Z_o^2}{Z_S}$, that is, $\frac{190^2}{500}$ or 72 ohms. Hence a complete match is effected, maximum transfer of energy results and no reflections occur.

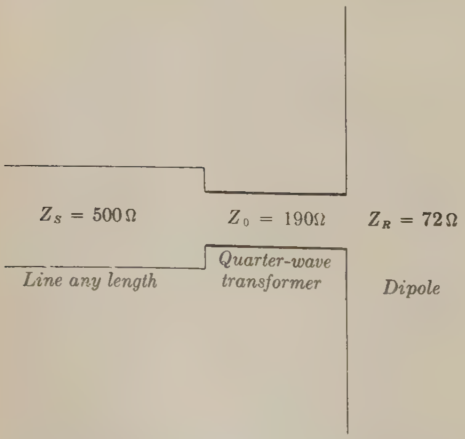


FIG. 10-59 A quarter-wave transformer used to match a 500Ω line to a 72Ω dipole.

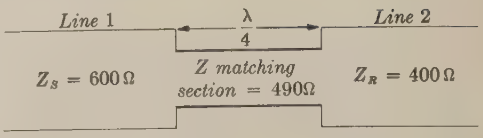


FIG. 10-60 A quarter-wave transformer used to match a 600Ω line to a 400Ω line.

Example 2. Find the characteristic impedance of a quarter-wave transformer required to match a 600Ω transmission line to a 400Ω transmission line (Fig. 10-60).

$$\begin{aligned} Z_o &= \sqrt{Z_S \times Z_R} \\ &= \sqrt{600 \times 400} \\ &\doteqdot 490 \end{aligned}$$

Therefore, the characteristic impedance of the matching section is 490Ω.

2. OPEN-ENDED HALF-WAVE SECTIONS.

Since a half-wave section repeats its load, that is, presents to its generator a

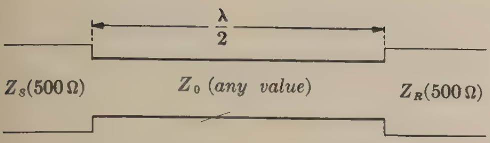


FIG. 10-61 A half-wave section used to connect a generator of impedance 500Ω to a load of equal impedance.

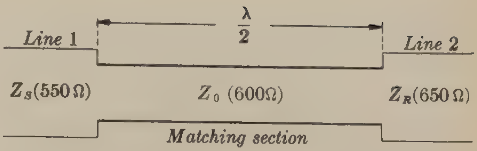


FIG. 10-62 A half-wave section used to connect a generator to a load of somewhat different impedance.

load equal to its termination regardless of the characteristic impedance, it can be used as a 1-to-1 transformer. Thus a half-wavelength section, or any multiple thereof, can be used to connect a generator to a load when their impedances are equal to each other, 500Ω each in Fig. 10-61. Further, if the impedances of the generator and the load do not differ too greatly from each other they may be connected by a half-wavelength line, or multiple thereof, whose characteristic impedance does not differ too greatly in value from the impedances to be connected. For example, line 1 (Fig. 10-62) with an impedance of 550 ohms may be connected to a load of 650 ohms by means of a 600-ohm half-wavelength section. The impedance match obtained in this case is suitable for most practical purposes.

F. Stub Lines

1. SINGLE-STUB TUNER.

A simple method of matching an aerial to the characteristic impedance of a transmission line in radar equipments using open-wire lines employs a short-circuited movable stub (Fig. 10-63). The stub line is connected across the main transmission line L_1 near the load end. Although both open-ended and short-circuited stubs may be used the latter is usually employed as it is stronger, easy to build and support and has lower losses.

To eliminate standing waves on the main transmission line L_1 the stub is placed across the line at points AB such that the line sees a load equal to its characteristic impedance Z_o. The impedance across AB is made up of the impedance of the stub line connected in parallel with the impedance of the line L_2. When such a match is effected, reflections and standing waves are eliminated from the main line between the generator and the stub. Since the standing waves on the part of the line L_2 between the stub and the load are not eliminated the stub must be located close to the load. The tapping points AB must be selected such that the impedance across AB has a resistive component equal to Z_o together with a reactive component. The reactive component is represented in the equivalent circuit of Fig. 10-64 by X_C and occurs even if the load impedance Z_R is a pure resistance since it is not matched

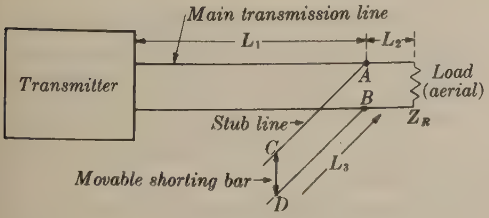


FIG. 10-63 A short-circuited movable stub used to match a line to an aerial.

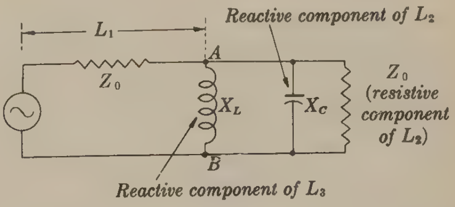

Reactive component of L_2
L_1
A
Z_0
X_L
X_C
Z_0 (resistive component of L_2)
B
Reactive component of L_3

FIG. 10-64 Equivalent circuit of the stub match of Fig. 10-63.

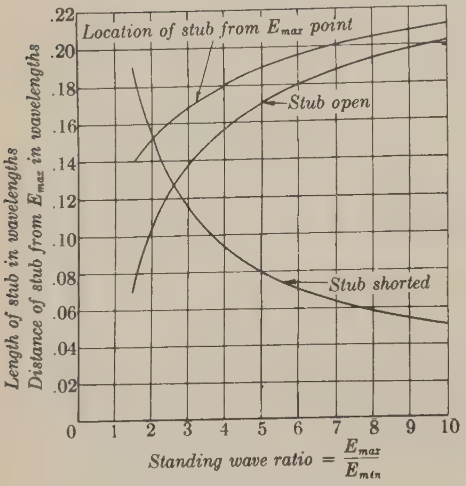


FIG. 10-65 Graph of stub match dimensions and standing-wave ratio.

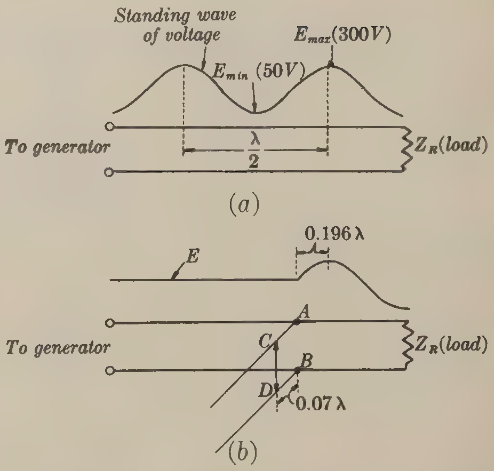


FIG. 10-66 Use of stub to match a line to its load (a) Standing wave of voltage before matching (b) Position and dimensions of stub and standing wave of voltage after match is set up.

to the impedance Z_o of L_1. The short-circuited stub which is usually made less than one quarter wavelength long must exhibit an impedance which is reactive and opposite in sign to the reactive component of L_2. Thus the impedance of the stub is represented in Fig. 10-64 by X_L. The position of the shorting bar CD on the stub is adjusted until $X_L = X_C$ and a pure resistance which is equal to Z_o results across AB and so the line L_1 is terminated in its characteristic impedance at AB.

To effect this match the location of the points A and B must be found and the correct adjustment of the shorting bar CD must be made. These two adjustments are difficult and tedious to make by the experimental method of cut and try. However, the approximate stub length and position can be obtained from the graph of Fig. 10-65 and then fine adjustments can be made to obtain optimum impedance match. To use the graph of Fig. 10-65 it is necessary to find the standing-wave ratio that exists on the line. This can be done by locating the voltage maximum nearest the load, that is, by means of an r-f voltage indicator and the voltage minimum a quarter-wavelength from this maximum point. Then, the standing-wave ratio is $\frac{E_{max}}{E_{min}}$. The stub length and position can then be determined directly from the graph. *Example.* Find the required stub length and position required to effect a match given $E_{max} = 300$ volts and $E_{min} = 50$ volts.

$$\text{Standing-wave ratio} = \frac{300}{50} = 6.$$

Then from the graph, the shorted stub length is 0.07λ and the position of the stub is 0.196λ from E_{max}.

The stub position is measured from the voltage maximum toward the generator for a short-circuited stub and from the voltage maximum toward the load for an open-circuited stub.

The relative location of the short-circuited stub in the above example is shown in Fig. 10-66 together with the standing waves of voltage which appear on the line before and after the stub is connected.

2. DOUBLE-STUB TUNER.

The single-stub tuner is suitable for open-wire lines but is not applicable to

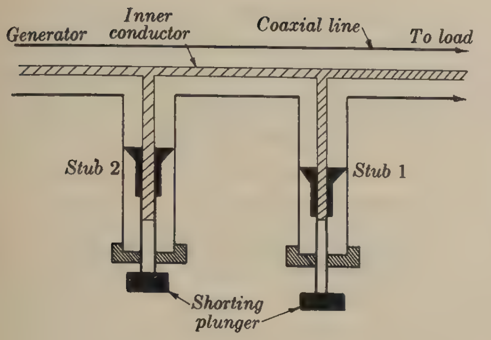


FIG.10-67 Double stub tuner to match a coaxial line to a generator.

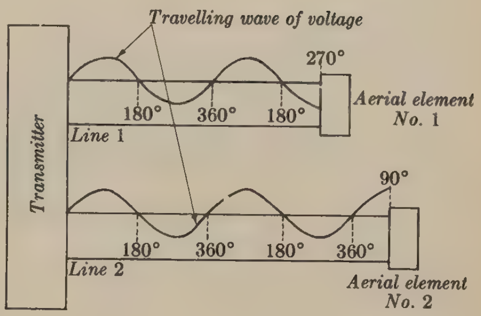


FIG.10-68 Use of non-resonant transmission line as a phase-shifting device to operate out-of-phase aerial elements.

coaxial lines because it is not practical to build a stub which can be moved along a coaxial line. However, the principle of stub matching can be applied to coaxial lines by using two stubs (Fig. 10-67) which are adjustable in length but fixed in position. This matching device is called a *double-stub tuner* and the stub lengths are adjusted by means of a movable shorting plunger. The movable plunger makes contact with both the inner and outer conductors of the coaxial stub. The separation of the stubs is sometimes one quarter wavelength but usually it is three eighths of a wavelength. The adjustment of the second stub adds inductance or capacitance to the line and so varies the position of the standing waves. This produces the same effect as if the first stub were moved along the line.

G. *Transmission Lines as Phase Shifters and Inverters*

1. PHASE SHIFTERS.

A non-resonant transmission line may be used as a phase-shifting device when it is necessary to operate two or more aerial elements with a phase difference between them. The phase difference is obtained by employing transmission lines of different lengths from the transmitter to the aerial elements as shown in Fig. 10-68. The diagram shows the travelling wave of voltage along the transmission lines and it is evident that there is a progressive phase shift per wavelength along each line. Thus the current in aerial 1 leads that in aerial 2 by 180° since line 2 is one half wavelength longer than line 1. This phase difference is dependable only in non-resonant lines, that is, only when no standing waves exist on the lines. Fig. 10-69 shows two aerial elements which form an aerial array. These elements are fed by two separate transmission lines, with the line feeding element 2 one quarter wavelength longer than the line-feeding element 1. Thus a 90° phase shift between the current in element 1 and that in element 2 is introduced. Similarly, pairs of aerial elements can be driven in the same phase by connecting the pairs at

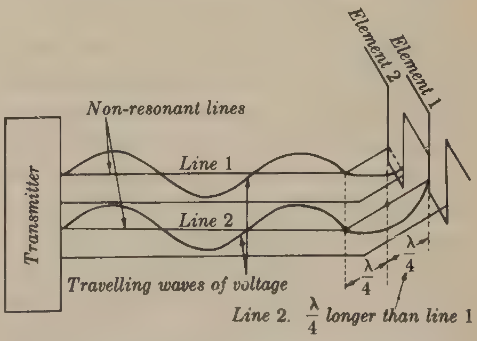


FIG.10-69 Non-resonant lines feeding a two-element aerial array.

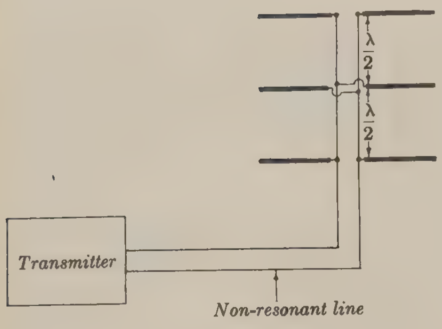


FIG. 10-70 Method of feeding pairs of aerial elements in phase by means of a non-resonant line.

intervals of $\frac{\lambda}{2}$ along a transmission line with reversed connections as shown in Fig. 10-70. At intervals of $\frac{\lambda}{2}$ along the line the voltage has opposite phases and the reverse connections serve to drive the elements in phase.

2. Phase Inverters.

It has been shown in section 10.13 E that a half-wavelength resonant-line section or any multiple thereof, can be used as a 1:1 transformer. In a transformer the secondary voltage is 180° out of phase with its primary voltage, Fig. 10-71 (a). Thus a half-wave section, or any odd multiple thereof, can be used to invert the voltage, the output voltage being 180° out of phase with the input voltage, Fig. 10-71 (b).

H. Coaxial-Line Balance Converters

In radar equipments it often becomes necessary to connect a coaxial line which is unbalanced to ground to a balanced line such as a shielded pair. Further, it is often desirable to transform the balanced line output from a transmitter which is obtained from a two-wire pick-up line to an unbalanced output so that a coaxial cable may be used as a transmission line to the aerial. A direct connection between unbalanced and balanced lines would produce a discontinuity in the electrical characteristics of the line. This condition would in turn produce reflections, a high standing-wave ratio and increased radiation loss. However, two such lines can be successfully connected by means of either a sleeve transformer or a trombone transformer.

1. The Sleeve Transformer.

The *sleeve transformer*, often referred to as a *bazooka* or *choke*, is simply a quarter-wave section and utilizes the principle that such a section inverts its load.

In the unbalanced coaxial line, the outer skin of the outside conductor is at ground potential, while the outer skin of the inner conductor is off ground potential and has a high impedance to ground. In the balanced two-wire line both conductors are off ground potential and exhibit a high impedance to ground. Thus, the sleeve transformer must cause the outer skin of the outside conductor of the coaxial line to exhibit a high impedance to ground at the point where it is to be connected to one wire of the balanced line.

To accomplish this, a quarter-wave

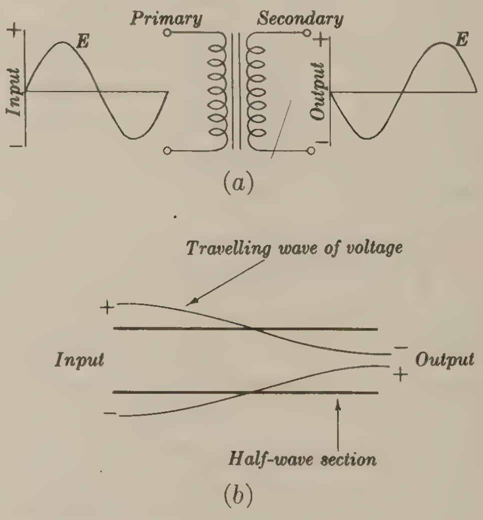


FIG. 10-71 Phase relationship between input and output voltages of a transformer and a half-wave section (a) Transformer (b) Half-wave section.

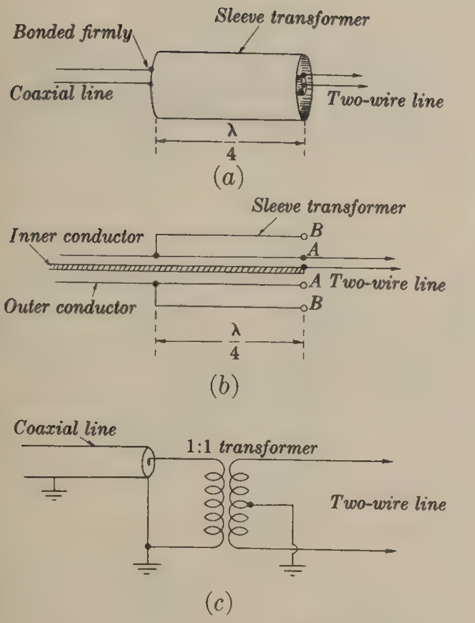


FIG. 10-72 Quarter-wave sleeve transformer (a) Construction (b) Method of connection (c) Equivalent circuit.

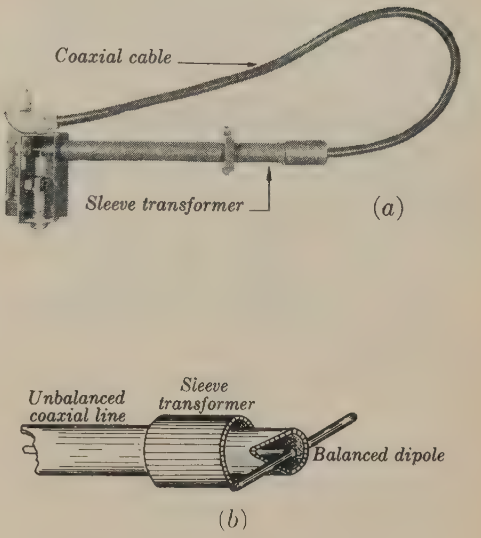


FIG. 10-73 Use of sleeve transformers (a) to connect a shielded two-wire line to a coaxial line (b) to transform an unbalanced line to a balanced line feeding a dipole.

shield is placed around the end of the coaxial line. This shield is connected and bonded firmly to the outer conductor of the coaxial line one quarter wavelength from the end of the line, Fig. 10-72 (a). Thus this shield and the outer conductor of the coaxial line form a quarter-wave section shorted at one end. Owing to the inverting action of a quarter-wave section there is a high impedance between *A* and *B*, Fig. 10-72 (b). Thus point *A* is isolated from ground and it can be connected to one wire of the balanced line. In this way, both wires of the two-wire line have a balanced high impedance to ground.

In effect the action of the sleeve transformer is similar to that of a 1:1 ratio transformer, with one primary lead grounded and with the secondary centre-tapped to ground. This equivalent circuit is shown in Fig. 10-72 (c). Typical sleeve transformers are shown in Fig. 10-73. Fig. 10-73 (a) shows a sleeve transformer which is used to connect a shielded two-wire line to a coaxial line. Fig. 10-73 (b) shows a typical sleeve transformer used to transform a coaxial line to a balanced line required to feed a 3 cm. dipole.

2. THE TROMBONE TRANSFORMER.

When a flexible coaxial cable is used to feed a dipole it is often most convenient to provide correct match of the unbalanced cable to the balanced aerial by means of a line balance converter made of the coaxial cable itself. Such a line balance converter is called a *trombone transformer*. Fig. 10-74 is a photograph of a flexible coaxial-cable trombone transformer which is used to transform the unbalanced cable to feed the balanced folded dipole of a Yagi array. Fig. 10-75 (a) and (b) show the physical and schematic diagrams illustrating how such a transformer may be used to convert from an unbalanced line to a balanced line.

The voltages of the two conductors in the balanced line, Fig. 10-75 (b), are

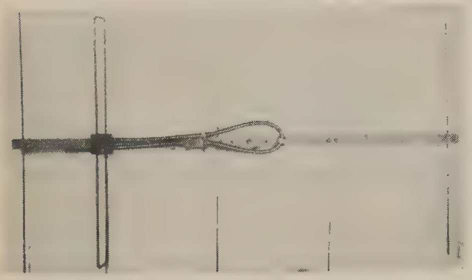


FIG. 10-74 A flexible coaxial-cable trombone transformer used to connect an unbalanced cable to the balanced folded dipole in a Yagi array.

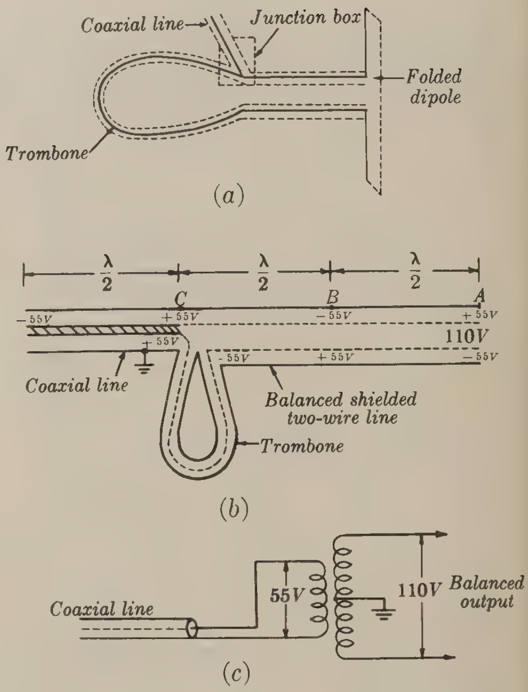


FIG. 10-75 Trombone transformer used to connect an unbalanced line to a balanced line (a) Method of connection (b) Detailed connection diagram (c) Equivalent circuit.

180° out of phase relative to ground as shown at *A*. A half-wavelength away at *B* the voltages have reversed, that is, the voltage along each conductor changes in phase by 180°.

At *C* one conductor is lengthened one half wavelength by means of the trombone transformer and so the voltage at the end of the lengthened conductor is in phase with that of the other conductor. Thus, the ends of the two conductors may now be connected together and an unbalanced line may be connected with its centre conductor joined to this junction. The currents of the two conductors of the balanced line add at *C* and flow together through the unbalanced-line centre conductor. The voltage supplied to the dipole by the balanced line is twice the voltage supplied to the balanced line by the unbalanced line. The transformation ratio is therefore 2:1 and the impedance ratio is 4:1; the equivalent circuit is shown in Fig. 10-75 (c). The output impedance of the trombone is usually matched to the impedance of the folded dipole by means of a quarter-wave section of proper impedance (Sec. 10.13 E) inserted between the output of the trombone and the folded dipole as shown in Fig. 10-75 (a).

I. Quarter-Wave Sections in Rotating Joints

When a radar aerial is to be rotated in azimuth, a rotating joint is provided in the aerial transmission line so that it is not necessary to rotate the entire set. In equipments employing open-wire transmission lines, an inductive coupling is often used. This coupling (Fig. 10-76) consists of two circular coils of one turn each mounted horizontally and parallel to each other. One coil turns with respect to the other without changing their mutual inductance appreciably. Each coil is shunted by a capacitor and

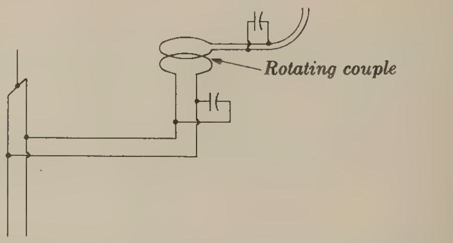


FIG. 10-76 Inductive coupling in a rotating joint in a transmission line.

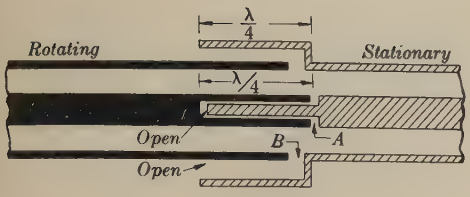


FIG.10-77 Quarter-wave section as a rotating joint in a coaxial line.

the coupling is in effect a 1:1 ratio transformer.

The rotating joint used in coaxial cables is shown in Fig. 10-77. To construct such a rotating joint the coaxial cable is cut and a cylindrical hole $\frac{\lambda}{4}$ in depth is drilled in the inner conductor on one side of the joint. A cylindrical $\frac{\lambda}{4}$ pin is attached to the inner conductor on the other side of the joint and it is of such size and length that it fits into the hole without touching the sides or bottom. The stationary outer conductor is fitted with a collar or concentric sleeve which surrounds the rotating outer conductor as shown. This collar is also $\frac{\lambda}{4}$ long and since this quarter-wave line section is open at its far end a low impedance is reflected at *B* and the outer conductor is thus made continuous for r-f. Similarly the inner conductor has a quarter-wave section added which is open at its far end and hence a low impedance is reflected at its input end *A*. Therefore the gap in the inner conductor at point *A* is in effect short circuited and this conductor is also made continuous for r-f.

The same type of rotating joint is often used when joining two sections of solid dielectric cable, such as telcothene cable. In this case the rotating joint consists of a short section of telescoped coaxial line made of brass tubing and rod with telcothene insulation. Usually for convenience, the lower half of the rotating joint bolts on to the bottom of an aerial junction box. The outer conductor projects upwards into the box where it fits into the outer tubing of the upper half of the rotating joint (Fig. 10-78). The centre of the lower half is the female portion of the inner conductor connection. The male portion is located in the upper half of the rotating joint.

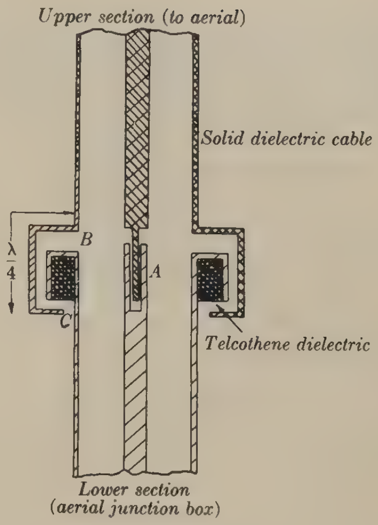


FIG.10-78 Quarter-wave section as a rotating joint in a solid-dielectric coaxial cable.

10.14 Introduction to waveguides. *Waveguides* represent a transitional stage between propagation of energy along wires and the radiation of energy in free space. *Guided waves* are those which follow a transmission line, while *unguided waves* are those which are radiated from an aerial. Radio waves are unguided electromagnetic waves which may be concentrated into a narrow beam by means of a directional aerial. Since the electromagnetic energy radiated in even a very narrow beam from a directional aerial is distributed over an increasing cross-sectional area as the distance from the aerial increases, attenuation results. However, energy in a guided wave is distributed over a relatively small, constant, cross-sectional area throughout the length of the transmission line regardless of bends and twists in the line. As a result almost all the energy put into the line reaches its destination.

Conventional theory regards the flow of current in a transmission line as a movement of electrons and the fields which exist between the wires as merely the result of this electron motion. However, it is equally possible to regard the fields as the fundamental quantities and the motion of the electrons as the result of the changing fields. The point of view which is adopted depends largely on the circumstance, the latter view becoming more useful at very-high frequencies. Because transmission line theory is more widely known than radiation theory, waveguides are often compared to transmission lines. This, however, leads into considerable complexity and in this text only the basis for comparison is given. Beyond this, the radiation theory is used to explain the operation of waveguides, because of its directness and relative simplicity. Thus in explaining microwave phenomena, the transmission of energy along a transmission line is regarded as taking place through the space between the conductors and not through the conductors themselves. The conductors guide the energy, which resides between the wires, in the form of waves from one point to another. Whereas it is convenient to explain transmission of energy via transmission lines in terms of the voltage and current distribution along the line it now becomes more useful to concentrate attention on the electric and magnetic fields of the wave. It is important, therefore, before proceeding to a study of waveguides to summarize the principal features of the electromagnetic fields of waves in free space and along transmission lines.

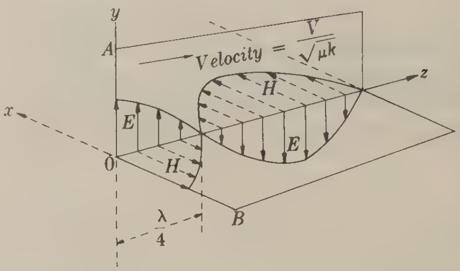


Fig. 10-79 Field distribution in a plane polarized electromagnetic wave.

10.15 Properties of electromagnetic waves in free space. Fig. 10-79 shows a sinusoidal plane-polarized electromagnetic wave travelling in free space. This wave is made up of oscillations of an electric field E and a magnetic field H in directions at right angles to each other and to the direction of propagation. The waves are transverse both in E and H and are, therefore, of the type TEM, transverse-electromagnetic; that is, there is no component of E or H in the direction of propagation. The velocity of propagation (v) in free space is equal to the velocity of light. In a medium of dielectric constant k and permeability μ,

$$\text{the velocity } = \frac{v}{\sqrt{K\mu}}.$$

When the medium is non-conducting E and H oscillate in phase with each other and the amplitudes of E and H are constant at all points in the wave front.

10.16 Properties of waves guided by transmission lines. The properties of waves guided by pairs of conductors are very similar to those of waves travelling in free space. Fig. 10-80 and Fig. 10-81 show the *field configuration* or *mode* formed by the lines of E and H for transmission along a coaxial cable and a shielded-pair line respectively. The electric and magnetic fields are represented by lines; the arrowheads indicate the direction of the fields while the line density shows the field intensities. The wave is propagated parallel to the axes of the conductors and is transverse both in E and H, that is, both the electric and magnetic fields are each perpen-

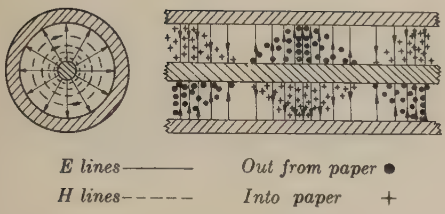


FIG.10-80 Pattern formed by the lines of E and H for transmission along a coaxial cable.

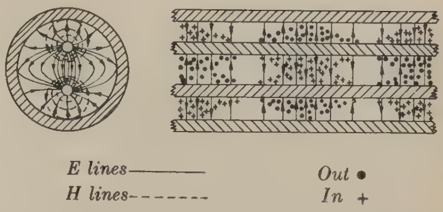


FIG.10-81 Pattern formed by the lines of E and H for transmission along a shielded pair.

dicular to the direction of travel and lie in a plane parallel to the wave front. The lines of magnetic force surround the inner conductors and form a pattern which is the same as that obtained by first considering a suitable steady current along the conductors but in opposite directions in each. At all points the two fields are mutually perpendicular. The velocity of the wave is the same as that for the unguided wave and is the same for all frequencies. There is no limit to the frequency of the wave which can be propagated. The mode shown in Fig. 10-80 and in Fig. 10-81 is the simplest one possible and is called the *dominant mode.* The coaxial line and the two-wire line have dominant modes and an infinite set of higher modes.

10.17 Types of waveguides. Any surface which separates distinctly two regions of different electric properties is capable of guiding electromagnetic waves. Such a surface may be one which separates a conductor from an insulator, or one which separates two insulators of widely different dielectric constants. As a result, the two-wire open or shielded line and coaxial cable are in effect waveguides. In addition, since even early radiation theory indicated that radio waves travel as do sound waves and since it is a well known fact that sound waves pass through pipes such as a voice tube by means of successive reflections from the sides, it is not at all surprising to find that electromagnetic waves pass through hollow metal pipes by successive reflections from the walls. The term *waveguide* is usually applied to this type of transmission line although other lines, strictly speaking, are also waveguides. Two different cross-sections of pipe, the rectangular waveguide and the circular waveguide are in common use. Fig. 10-82 shows a rectangular waveguide, the cross-section indicating a and b dimensions.

10.18 Advantages of hollow waveguides. A hollow waveguide has lower losses than an open two-wire line or coaxial line in the frequency ranges practical for each. The open two-wire transmission line has three kinds of losses

(i) radiation loss
(ii) dielectric loss
(iii) copper loss.

The coaxial line has negligible radiation loss since the outer conductor acts as a shield which confines the fields within the space between inner and outer conductors. The hollow waveguide also has negligible radiation loss for the same reason.

In coaxial cable the dielectric loss in

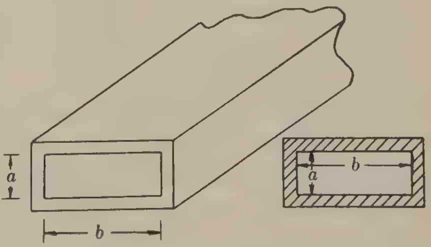


FIG.10-82 Rectangular waveguide and cross-section showing a and b dimensions.

the solid insulation or in polystyrene beads is appreciable at very high frequencies. However air has negligible dielectric loss at any frequency. As a result hollow waveguides, which have air dielectric, have negligible dielectric loss.

In a transmission line the copper loss increases as the frequency increases since at high frequencies the current flows in a thin layer near the surface of the conductor. As the frequency increases, the thickness of the conducting layer decreases which causes the copper loss to increase as the effective resistance of the conductor increases. The currents responsible for the copper loss are equal in magnitude in the inner and outer conductors, but since they are forced by the skin effect to flow in the surface layers of these conductors they dissipate most of their energy in the inner conductor which has the smaller surface. In a hollow waveguide the inner conductor is eliminated and hence the copper loss is greatly reduced. A coaxial cable with its inner conductor removed becomes a circular hollow waveguide. The copper loss of a hollow waveguide, rectangular or circular, is less than that of a coaxial cable of the same size operated at the same frequency.

Thus the hollow waveguide has small copper loss and for all practical purposes neither radiation loss nor dielectric loss. Hence hollow waveguides are more efficient than other lines and the attenuation in them is considerably less. Loss of energy from a wave as it travels along a transmission line is objectionable not only because it attenuates the signal which arrives at the other end, but also because of the undesirable heating produced by the wasted energy. In the case of high-power transmission this heating is often sufficient to soften the dielectric in the cable and cause breakdown.

A second advantage of a waveguide is that it is capable of transmitting higher power than a coaxial cable of the same size. The power transmitted on a coaxial cable is given by

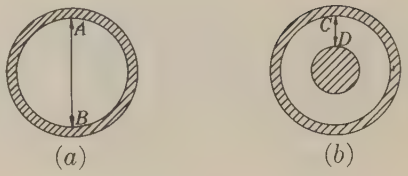


FIG. 10-83 Comparison of length of dielectric breakdown paths in (a) Waveguide (b) Coaxial cable.

$$P = \frac{E^2}{Z_o}$$

where P = the power
E = the voltage
Z_o = the characteristic impedance of the line.

Hence, since Z_o is constant, the power may only be increased by increasing the voltage. This is limited by the breakdown voltage of the dielectric between the conductors. In a circular waveguide, Fig. 10-83 (a), the maximum voltage E appears across AB while in the coaxial cable it appears across CD, Fig. 10-83 (b). Since AB is greater than CD the voltage at which breakdown occurs in a waveguide is greater than that at which breakdown occurs in a coaxial cable. As a result, the power-handling capability of a waveguide is higher than that of a coaxial cable of the same size.

A third advantage of a waveguide is its simple construction since insulated supports for the inner conductor as in other lines are not required. In addition, the waveguide is less subject to mechanical injury due to its rugged nature.

10.19 Development of a waveguide from an open two-wire transmission line. The use of quarter-wave stubs as insulators for transmission lines at high frequencies has been discussed in section 10.13B. Such insulators are useful because of the low losses which they introduce, but they are restricted to very

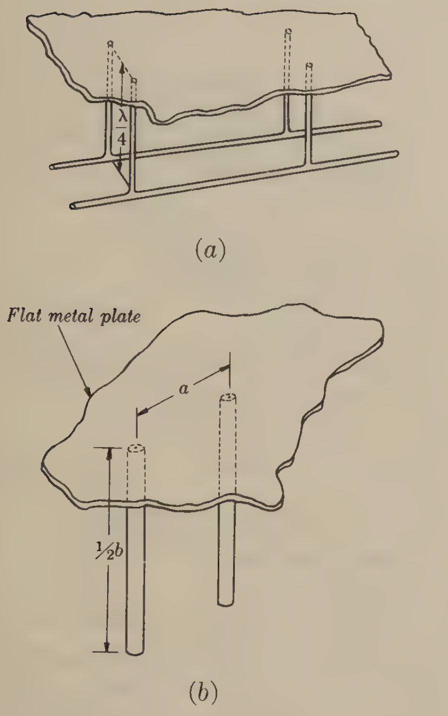


FIG.10-84 Open two-wire line supported by quarter-wave stub insulators (a) Construction (b) Enlarged view of an insulator.

high frequencies because of their physical size. For example, to make such an insulator for a 60-cycle power line it would have to be approximately 800 miles long. An open two-wire transmission line supported by such quarter-wave stub insulators is shown in Fig. 10-84 (a). Such a line may be used for only one frequency or for a very narrow band of frequencies, since the stub length is only a quarter wavelength at the particular frequency for which it was cut. Actually in measuring off the quarter wavelength of a stub as shown in Fig. 10-85 the distance $\frac{1}{2}b+\frac{1}{2}a$ should be approximately $\frac{\lambda}{4}$ provided $\frac{1}{2}a$ is small compared to $\frac{1}{2}b$. If a is large compared to b the impedance per unit length along the two sections may be sufficiently different to cause an appreciable error. However, this situation does not occur in the construction shown in Fig. 10-84 (b). In this case the stub lines are fastened to a large flat metal plate. Thus very few magnetic lines of force encircle this plate in the distance a. Therefore the dimension a is not critical and may be neglected in measuring the length of the stub. That is, the length $\frac{1}{2}b$ is made about one quarter wavelength.

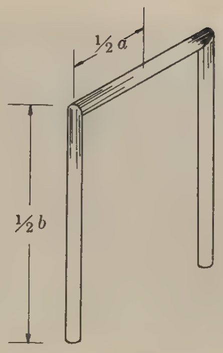


FIG.10-85 Dimensions of a quarter-wave stub insulator.

Fig. 10-86 shows a two-wire line supported from both top and bottom by quarter-wave sections thus making half-wave rectangular frames. If the number of these stubs is increased without limit until they form a rectangular pipe, a waveguide is produced (Fig. 10-87). A waveguide can be thought of as being made up of two bus-bars and a multitude

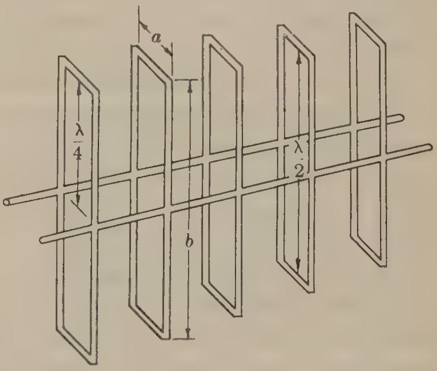
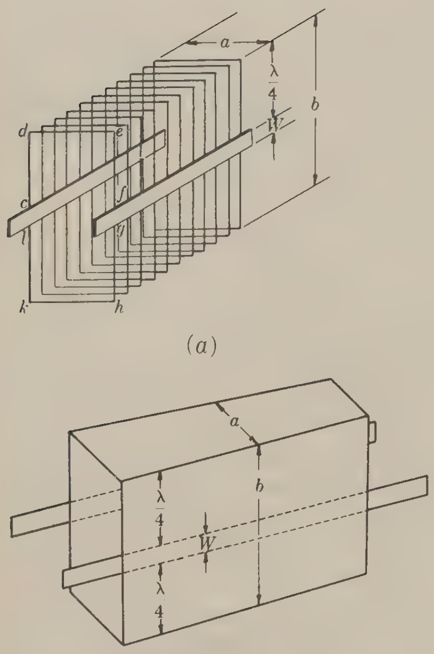


FIG.10-86 Two-wire line supported by quarter-wave stubs on top and bottom forming half-wavelength, rectangular, supporting frames.



(a)

(b)

FIG. 10-87 Two bus-bars supported by many $\frac{\lambda}{4}$ stub insulators to form a waveguide.

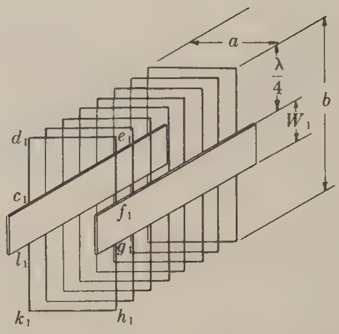


FIG. 10-88 Diagram to aid the explanation of why a waveguide transmits all frequencies higher than a limiting minimum frequency.

of quarter-wave stub insulators or half-wave rectangular frames.

A waveguide has the advantage of being able to carry a relatively large band of frequencies. It can carry not only a fundamental or limiting frequency but all higher frequencies. The reason for this can be best understood by a consideration of Figs. 10-87 and 10-88. In Fig. 10-87 the waveguide consists, in effect, of two bus-bars supported by upper and lower quarter-wave stub insulators. The distances of *cd*, *ef*, *gh*, *kl* equal $\frac{\lambda}{4}$ and distance W is equal to the theoretical width of the bus-bar. At some higher frequency Fig. 10-88 illustrates the conditions that exist, the dimensions a and b of the waveguide being kept constant. The width of the bus-bar may be considered to increase to W_1, while the quarter-wave stub insulators decrease in length with distances c_1d_1, e_1f_1, g_1h_1, k_1l_1 equal $\frac{\lambda}{4}$ at the new frequency.

Similarly at some lower frequency (longer wavelength) the bus-bar may be considered to decrease to some new width while the quarter-wave stubs increase in length. At a certain low frequency the bus-bar width becomes zero where the two quarter-wave stubs together equal b. For all frequencies lower than this the two stubs are less than $\frac{\lambda}{4}$. Therefore instead of providing the desired high resistive impedance they present a much lower inductive impedance and the current is shorted out stopping transmission. At the frequency for which b is exactly one half wavelength, the wave reflects back and forth across the guide, making no progress down the guide. Thus a waveguide has a *critical minimum frequency*, called the *cut-off frequency*, below which it will not transmit energy.

The value of the cut-off frequency is, as indicated above, closely related to the dimensions of the waveguide. However, when the line becomes a waveguide (Fig. 10-87) by increasing the number of stubs, the lines of magnetic force can no longer link the yoke portion a of the stubs. Hence the dimension a is not critical in determining the cut-off frequency of a waveguide. However, this dimension *a* determines the breakdown voltage of the guide so that for high-power transfer it must be large. In addition it is easier mechanically to build *a* large. In practice *a* is usually made about one-half of *b*.

For a rectangular waveguide there are simple relationships between cut-off frequency and the dimensions. For example for one particular mode of transmission the relationship is

$$\lambda_c = 2b \text{ or } \frac{v}{f_c} = 2b$$

where

λ_c = maximum free-space wavelength
b = inside width of pipe
v = velocity of wave propagation in free space (3×10^8 m/s)
f_c = cut-off frequency.

In construction of a rectangular waveguide for 10-cm. application, *b* would be made slightly wider than 5 cm., in practice, about 7.0 cm. ($2\frac{3}{4}$ in.) and dimension *a* about 3.5 cm. ($1\frac{2}{5}$ in.). For 3-cm. applications, a waveguide with outside dimensions of 1 in. × ½ in. is commonly used. For wavelengths longer than 10 cm. the sizes of the pipes are impractical.

Of course a waveguide is not built by welding upper and lower quarter-wave stubs together. Any such method would be highly impractical. In practice a waveguide is constructed as a simple piece of rectangular pipe of the required dimensions as shown in Fig. 10-89.

10.20 Electric and magnetic fields in waveguides. In order to understand how a waveguide is excited and how energy is removed at its termination it is important to know the distribution of fields inside the guide.

A. Electric Fields

Fig. 10-90 shows a section of two-wire transmission line shorted at both ends and cut to $\frac{3}{2}\lambda$ at some frequency. Consider this line to be supported by upper and lower quarter-wave stubs, each pair of upper and lower stubs making up a rectangular frame. As a result of the shorted ends when the line is energized, reflections occur producing standing waves of current and voltage both in the line itself and on the stubs. Fig. 10-90 shows only the *E* lines. These lines represent the electric field between the conductors. The arrowheads indicate the direction of the field while the line density shows the field intensity. For simplicity, the distribution is considered for only the instant in the cycle when the voltages are greatest. The maximum standing-wave voltages occur at *BB*, *DD'*, and *FF'*. Since the various insulating frames are excited to resonance by the line, they have as their maximum the voltage across the line at the point where the frames are attached. The two frames drawn separately below the line show that the maximum voltage is different at different positions along the line. The *E* lines are always perpendicular to the conducting surfaces.

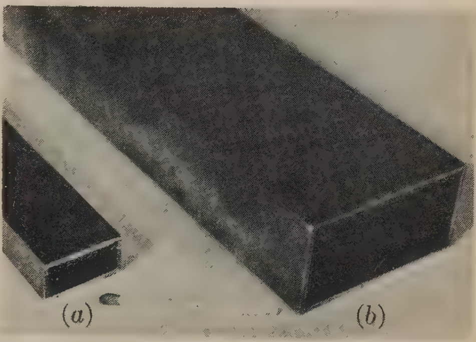


FIG. 10-89 Rectangular waveguides (a) 3 cm. (b) 10 cm.

If the line of Fig. 10-90 is formed into a waveguide by increasing the number of frames the electric field in such a waveguide has the distribution shown in Fig. 10-91. The field distribution on the quarter-wave stubs combines with that on the line itself to form a field of

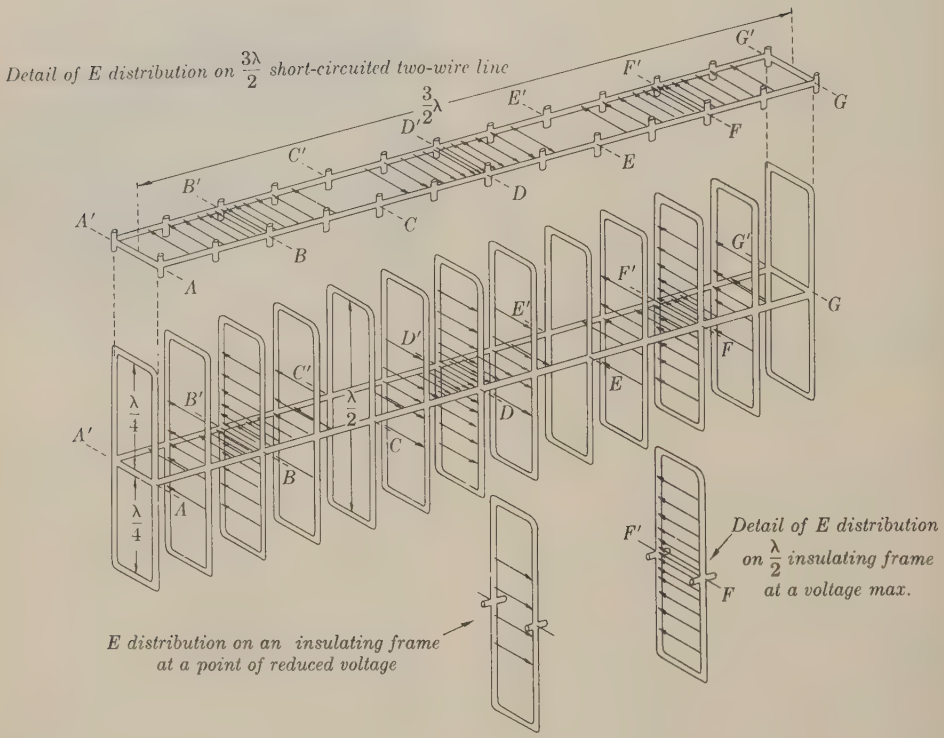

Detail of E distribution on $\frac{3\lambda}{2}$ short-circuited two-wire line
$\frac{3}{2}\lambda$
A′
A
B′
B
C′
C
D′
D
E′
E
F′
F
G′
G
$\frac{\lambda}{4}$
$\frac{\lambda}{4}$
$\frac{\lambda}{2}$
Detail of E distribution on $\frac{\lambda}{2}$ insulating frame at a voltage max.
E distribution on an insulating frame at a point of reduced voltage

FIG. 10-90 Distribution of E lines on an energized $\frac{3}{2}\lambda$ short-circuited two-wire line supported by pairs of upper and lower $\frac{\lambda}{4}$ stubs.

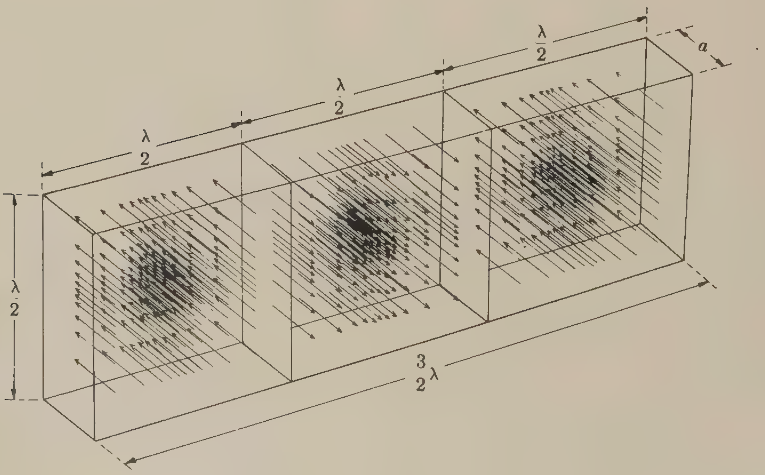


FIG. 10-91 E line distribution inside a $\frac{3}{2}\lambda$ section of short-circuited energized waveguide.

high intensity in the centre of each half-wave guide section. The E lines exist in a standing wave similar to the standing wave of voltage which exists on a shorted line. Thus the E lines build up to a maximum, decrease to zero, build up to a maximum in the opposite direction and decrease again to zero. The lines maintain a constant maximum density at any one location. Of course, an r-f voltage exists between the sides of the waveguide on which the E lines terminate. In the centre of each half-wave guide section the voltage across the guide is a maximum since the field is a maximum. The distance a must be large enough to keep this voltage from causing arc-overs in the guide.

B. *Magnetic Field*

To understand the magnetic field distribution in a waveguide it is well to recall first the magnetic field distribution about a current-carrying solenoid. Fig. 10-92 (a) shows the magnetic field about a current-carrying loop with the lines of force encircling the conductor. If many such loops are placed close together to form the solenoid of Fig. 10-92 (b) the lines of force due to each loop add to produce a strong magnetic field. The H lines form loops being closed outside the solenoid as shown in the diagram.

Similarly, if Fig. 10-93 represents a section of a two-wire transmission line supported by upper and lower quarter-wave stubs or frames the H lines about the line and the frames are as shown in the diagram. If the frames are relatively close together the H lines cancel between conductors and add inside the frames to produce an H loop entirely inside the frame as shown in Fig. 10-94.

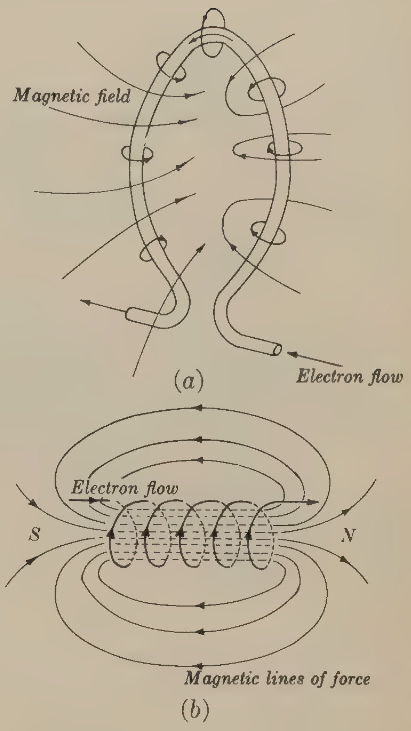


FIG.10-92 Magnetic field about a current-carrying loop and solenoid (a) Loop (b) Solenoid.

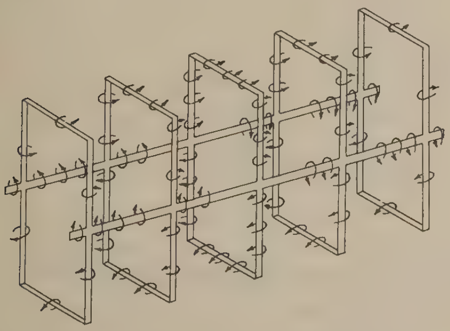


FIG.10-93 H lines about an energized two-wire line supported by upper and lower $\frac{\lambda}{4}$ stubs.

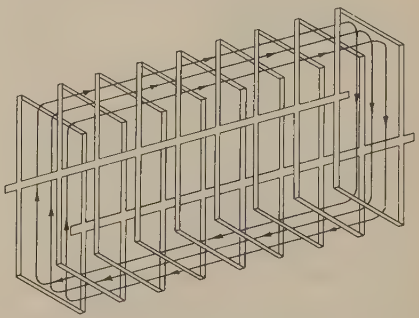


FIG.10-94 H lines forming complete loop lines inside the frames of a two-wire line with $\frac{\lambda}{4}$ stubs close together.

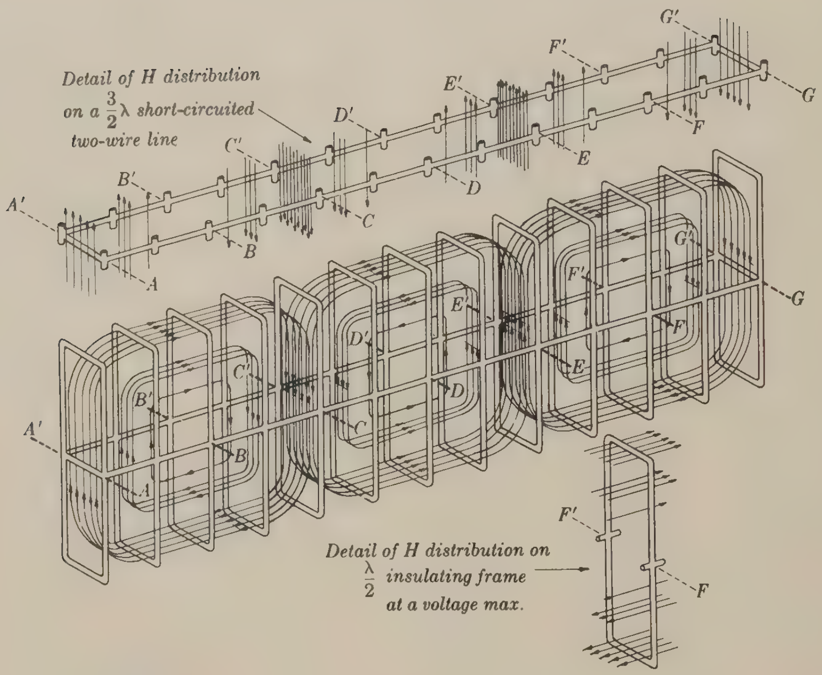

Detail of H distribution on a $\frac{3}{2}\lambda$ short-circuited two-wire line
A′
A
B′
B
C′
C
D′
D
E′
E
F′
F
G′
G
Detail of H distribution on $\frac{\lambda}{2}$ insulating frame at a voltage max.

Fig. 10-95 H lines in an energized $\frac{3}{2}\lambda$ short-circuited two-wire line supported by pairs of upper and lower $\frac{\lambda}{4}$ stubs.

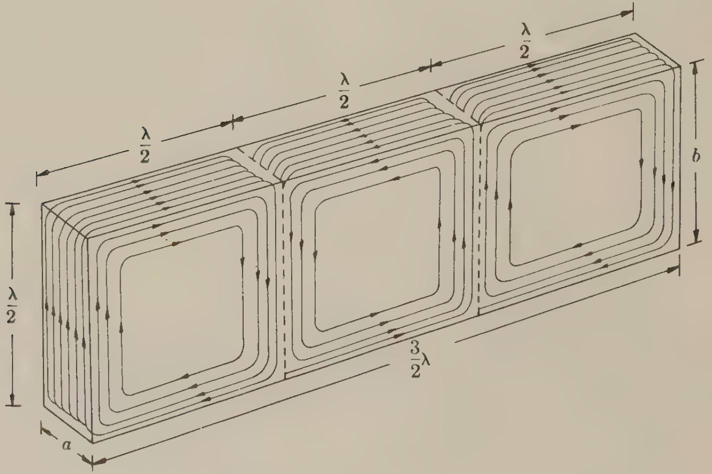


Fig. 10-96 H line distribution inside a $\frac{3}{2}\lambda$ section of short-circuited energized waveguide.

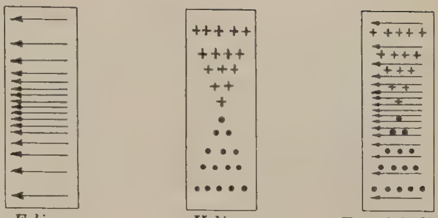
Thus, Fig. 10-95 shows the magnetic field distribution about a short-circuited section of a two-wire line cut to $\frac{3}{2}\lambda$ at some particular frequency. This line is supported by upper and lower $\frac{\lambda}{4}$ stubs and is energized by a generator. The H lines form complete loops entirely within the frames of each $\frac{\lambda}{2}$ section of line. The arrowheads show the direction of the field at the instant represented in the diagram. The density of the lines shows the intensity of the magnetic field. The electric field which is, of course, also present is not shown to avoid complexity. The frame shown below the main section of the diagram indicates the detail in the frame which appears at a voltage maximum. The *H* lines are always parallel to the conducting surface.

If the transmission line of Fig. 10-95 is developed into a waveguide by increasing the number of frames without limit the *H* line distribution is as shown in Fig. 10-96. These lines appear, as in Fig. 10-95, as complete loops spaced one half wavelength apart. There are no *H* lines outside the guide as long as it is completely closed.

C. *Diagrammatic Representation of both E and H Lines in a Waveguide*

Since both the electric field and the magnetic field are actually inseparable and always exist at the same time in a waveguide it becomes necessary to indicate both these fields on a diagram in a way which does not make the diagram confusing. It is evident that combining Figs. 10-91 and 10-96 into a single diagram would result in too complex a diagram for understanding. As a result cross-section diagrams of a waveguide are usually employed.

Thus if a cross-section of a waveguide in Fig. 10-91 is taken halfway between any $\frac{\lambda}{2}$ section, the *E* line distribution is as shown in Fig. 10-97 (a). Also, if a similar cross-section is taken of Fig. 10-96 the *H* line distribution is as shown in Fig. 10-97 (b). The combination of these two diagrams in Fig. 10-97 (c) shows both the *E* and *H* line distributions.



E lines (*a*)

H lines (*b*)

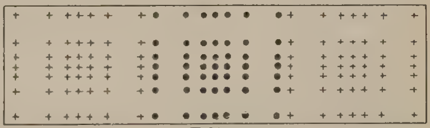
E and H lines (*c*)

+ *Into paper*
• *Out from paper*

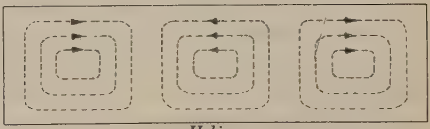
Fig. 10-97 Cross-sectional end view of waveguide showing typical field distributions (a) Electric field (b) Magnetic field (c) Electromagnetic field.

A second method of showing both the *E* and *H* lines is to present a side view of the waveguide. Fig. 10-98 (a) shows a side view of Fig. 10-91; Fig. 10-98 (b) shows a side view of Fig. 10-96. The combined side view with both *E* and *H* line distribution is shown in Fig. 10-98 (c).

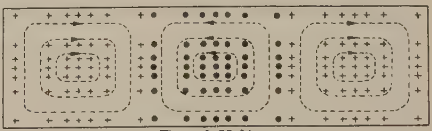
It should be noted that when a waveguide is used as a transmission line it should not be operated with standing waves. Like any other form of trans-



E lines
(*a*)



H lines
(*b*)



E and H lines
(*c*)

+ *Into paper*
• *Out from paper*

Fig. 10-98 Cross-sectional side view of waveguide showing field distributions (a) Electric field (b) Magnetic field (c) Electromagnetic field.

mission line it may be operated without standing waves when properly terminated. Such operation is necessary to secure maximum transfer of energy.

10.21 Modes of transmission in waveguides. The distribution of E and H lines inside a waveguide can take many forms depending upon the manner in which the waveguide is excited. Each of these forms is called a *mode of operation.* To differentiate between these modes a system of numbering has been devised. In all of the modes either the magnetic or the electric field must be perpendicular to the direction of propagation of energy down the guide. Hence any mode must be either *transverse electric* (TE) or *transverse magnetic* (TM). When a guide is excited to operate in the TE mode the E lines are across the guide and no E lines lie lengthwise along the guide; similarly, in a TM mode the H lines form loops in planes perpendicular to the walls of the guide and no part of an H line lies lengthwise along the guide.

To complete the description of the mode, subscript numbers are used with the letters TE or TM. For rectangular guides the subscript numbers indicate the following characteristics:

(i) 1st subscript number indicates the number of half-wave patterns of the transverse lines which exist along the short dimension a of the guide through the centre of the cross-section.

(ii) 2nd subscript number indicates the number of transverse half-wave patterns that exist along the long dimension b of the guide through the centre of cross-section.

If there is no pattern along one dimension of the guide a zero is used.

For example, the use of this nomenclature may be illustrated by applying it to the mode of operation indicated by the field distribution shown in Fig. 10-97 and 10-98. Since the electric field (Fig. 10-97) is perpendicular to the longer wall of the guide the electric field is transverse and so the mode must be a TE mode. In order to determine the subscript numbers it is only necessary to examine Fig. 10-97 (c). Since there is no change in the electric field intensity along the short dimension through the centre, that is, there is no half-wave pattern, the first subscript is zero and the mode is a TE_0 mode. However, along the long dimension through the centre the electric field intensity goes from zero at the bottom through a maximum at the centre and back to zero at the top of the cross-section. Thus the field distribution along the vertical through the centre is of the form of a sine wave in which one half-wave pattern exists in this direction. Thus this mode becomes the $TE_{0,1}$ mode. This is the mode most often used in practice. In this mode since there is no electric field in the direction of propagation and since there is a magnetic field in the direction of propagation this mode is sometimes referred to as a *magnetic wave.*

In applying this nomenclature to circular waveguides the rule for the letters TE or TM is as for the rectangular guide. However, in this case, the first subscript indicates the number of whole (or full-wave) patterns of the lines encountered around the circumference of the guide. The second number indicates the number of half-wave patterns that exist along a diameter.

Four modes of transmission in rectangular waveguides and two modes of transmission in circular waveguides are shown in Fig. 10-99. The diagram indicates the number of half-wave and full-wave patterns for each mode and thus the nomenclature shown in each case may be verified. In addition the formula for determining the cut-off wavelength (λ_c) that can be transmitted by each waveguide when operated in the mode

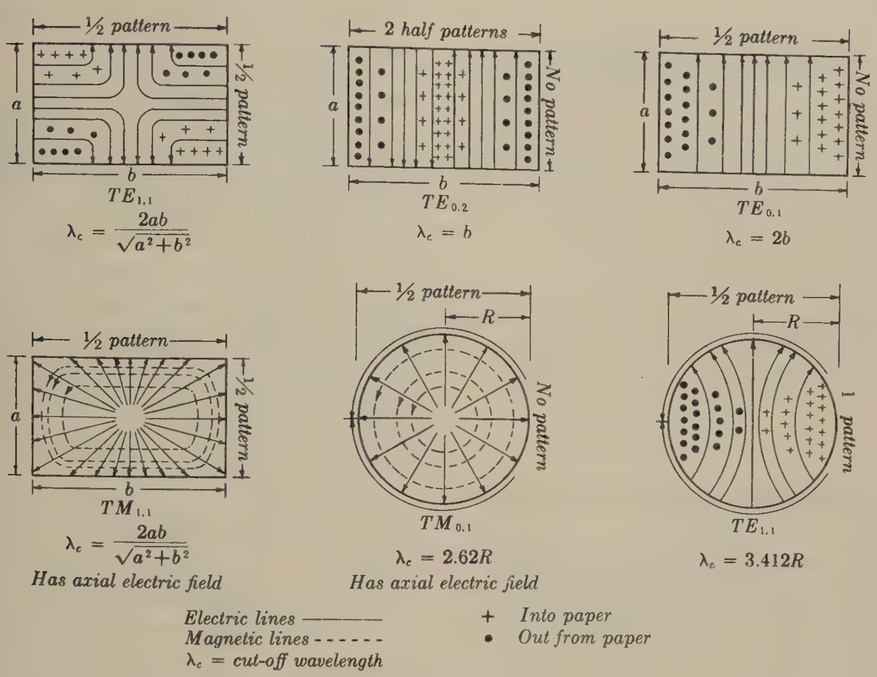


FIG. 10-99 End views of six modes of transmission in waveguides with formulas for determining cut-off wavelength.

shown is given. The $TM_{1,1}$ and $TM_{0,1}$ modes are transverse magnetic in spite of the fact that the E lines appear to be transverse. In both these modes part of the E lines lie along the axis of the guide, that is, in the direction of transmission down the guide.

The manner in which a waveguide may be developed from a two-wire transmission line to operate in a $TE_{0,2}$ mode is shown in Fig. 10-100. The transmission line in this case may be considered to be supported from above by $\frac{\lambda}{4}$ stubs and from below by $\frac{3}{4}\lambda$ stubs. This arrangement can be developed into a complete waveguide by increasing the number of insulating supporting frames to form a surface. This waveguide may also be considered to be made up of two two-wire transmission lines, one at LL' and the other at MM'. These lines are now separated by a half-wavelength section LM and each line is supported by $\frac{\lambda}{4}$ stubs. Thus this arrangement is in effect the same arrangement for two $TE_{0,1}$ modes acting together at a phase angle of 180°. The field distribution is shown in the side view (Fig. 10-101) and by applying the nomenclature for naming modes of transmission in waveguides

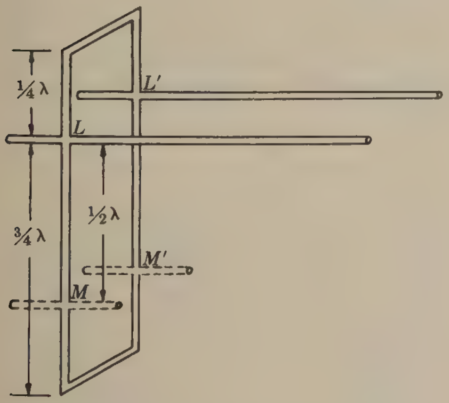


FIG. 10-100 Development of a waveguide from a two-wire transmission line to operate in the $TE_{0,2}$ mode.

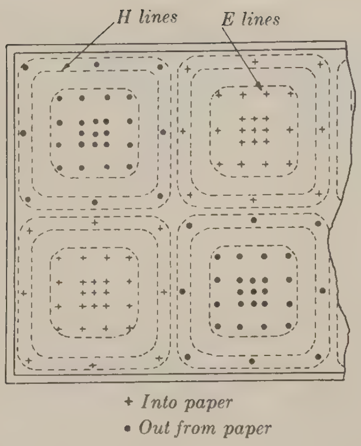


FIG. 10-101 Side view of field distribution in a $TE_{0,2}$ mode.

the mode is the $TE_{0,2}$. In this waveguide the cut-off frequency is doubled; however, this is of no particular advantage as it now takes twice as wide a guide as that for the $TE_{0,1}$ mode.

Circular waveguides may be developed from two-wire transmission lines by using two quarter-wave semi-circular stubs as shown in Fig. 10-102. The field distribution in such a guide is shown in Fig. 10-103 and the mode of operation is $TE_{1,1}$. The maximum wavelength that can be transmitted is given by the formula

$$\lambda_c = 1.7 \times d$$

where d = the diameter
λ_c = the cut-off wavelength.

This mode for a circular waveguide is almost the same as the $TE_{0,1}$ mode for a rectangular guide.

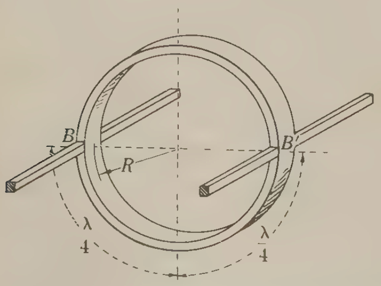


FIG. 10-102 Development of circular waveguide from a two-wire transmission line.

10.22 Transmission of energy through a waveguide. In the development of waveguides thus far, only a $\frac{3}{2}\lambda$ section closed at each end has been considered with standing waves present both lengthwise and crosswise in the guide. Hence in Fig. 10-98 the maximum density of E lines is 90° from the maximum density of H lines since the energy is reflecting back and forth across the guide and is not travelling down the guide. This relationship is the same as that which exists between voltage and current in a shorted transmission line.

However, in a very long waveguide with a means of injecting the energy and terminating it so that no reflections are produced, the standing waves disappear in the direction of travel. This is similar to the condition which exists on a long two-wire transmission line terminated in its characteristic impedance. In this case voltage and current are in phase. Similarly in the waveguide, E and H are in phase and this can be represented diagrammatically by showing the E maxima coinciding with the H maxima.

Fig. 10-104 shows the two most important modes $TE_{0,1}$ and $TM_{0,1}$ used for transmission in circular waveguides together with the most important mode $TE_{0,1}$ used in rectangular guides. In these diagrams maximum values of E and H occur simultaneously and energy is assumed to be transmitted down the guide.

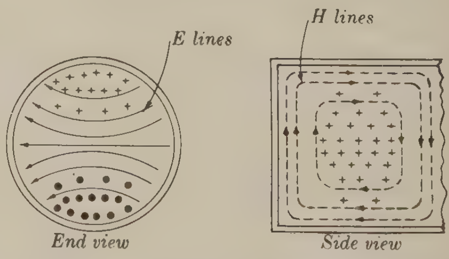


FIG. 10-103 End view and side view of field distribution in circular waveguide operated in $TE_{1,1}$ mode.

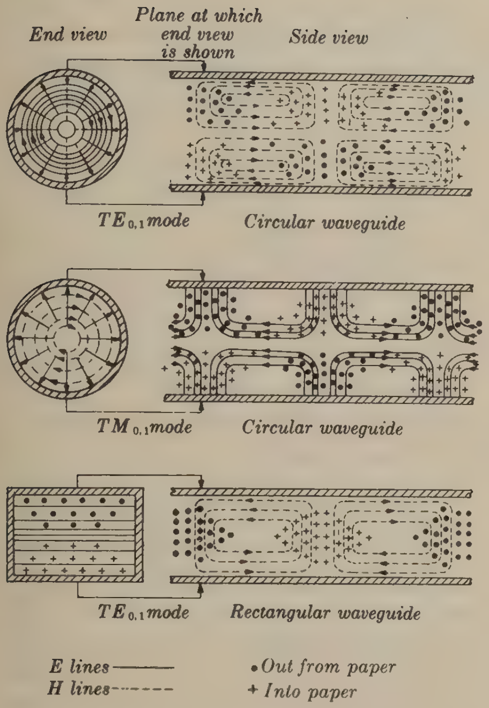


FIG. 10-104 The two most useful modes $TE_{0,1}$ and $TM_{0,1}$ for transmission in circular guides and the most useful mode $TE_{0,1}$ for rectangular guides.

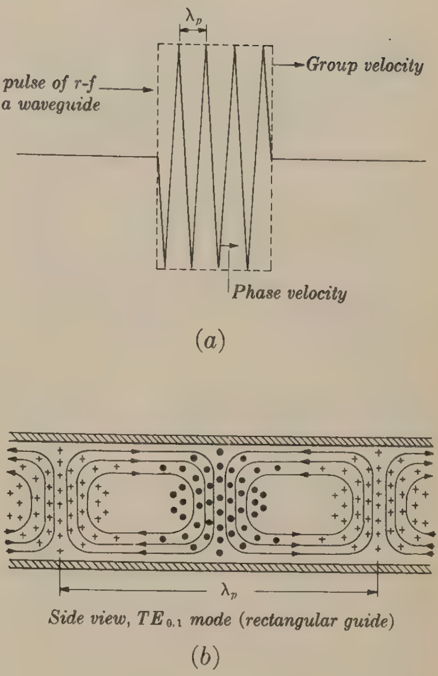

λ_p
Group velocity
A pulse of r-f in a waveguide
Phase velocity
(a)
λ_p
Side view, $TE_{0,1}$ mode (rectangular guide)
(b)

FIG. 10-105 Phase and group velocities in a waveguide (a) Relationship between phase and group velocities for a pulse transmission (b) Wavelength corresponding to phase velocity.

The velocity of electromagnetic waves in the guide is actually considerably less than the velocity in free space since the waves travelling with the velocity of light progress by successive reflections at the sides of the guide and hence must travel farther than the length of the guide. The velocity of propagation of energy down the guide is called the *group velocity* and is less than the velocity of light. The meaning of group velocity is further understood by a consideration of Fig. 105 (a) which shows the waveform of an r-f pulse of electromagnetic radiation. Consider this pulse to be propagated down a waveguide. The pulse takes a certain time to travel from the input end to the output end. The velocity at which the pulse travels down through the guide is the group velocity. In addition, in a waveguide, the r-f cycles move within the pulse envelope in the direction of travel of the pulse. The velocity at which the cycles move is greater than the group velocity and is called the *phase velocity.* This superimposed velocity which is the change of field intensity or the velocity of the wave crest along the side of the guide is also greater than the velocity of light. The wavelength λ_p corresponding to this phase velocity is illustrated in Fig. 10-105 (b). If the waves travelling with the velocity of light reflect at the sides of the guide at angles approaching 90° the phase velocity may approach infinity. The wavelength λ_p also increases and may approach infinity. Under this condition the wave no longer progresses down the guide. That is, the group velocity is zero and the r-f is at the cut-off frequency. It is important to keep in mind that the apparent wavelength λ_p for a practical waveguide may

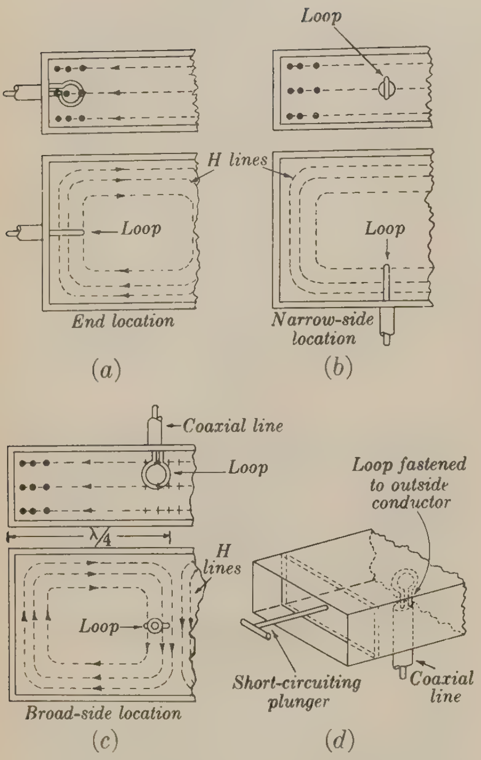


Fig. 10-106 Inductive coupling between a coaxial line and a waveguide (a) End location of loop (b) Narrow-side location of loop (c) Broad-side location of loop (d) Short-circuiting plunger as matching device.

be 1.5 to 2.0 times the wavelength in air and this is the value which must be used in all calculations.

10.23 Methods of energizing and extracting energy from a waveguide. There are two principle methods in use by which energy is put into, or removed from, a waveguide.

(i) Inductive or loop coupling
(ii) Electric or probe coupling.

A. Inductive or Loop Coupling

Fig. 10-106 shows three methods of inductive coupling from a coaxial cable to a waveguide. A coupling loop is placed in the guide so that it is always located where the *H* field intensity is at a maximum when energy is being taken out from the guide, or where maximum *H* field intensity is desired when energy is being put into the guide. The plane of the loop is oriented perpendicular to the direction of the lines of magnetic force. Coupling may be varied by rotating the loop, moving it from a position of maximum flux, changing its size or shielding it. The length of the waveguide is not critical but the distance from the loop to the near end of the guide is important. As a result a *short-circuiting plunger* or *piston*, Fig. 10-106 (d), the position of which can be adjusted to cause the short length of the guide between it and the loop to act as a matching device, is used.

B. Electric or Probe Coupling

The distribution of the magnetic and electric fields about a quarter-wave Marconi aerial is as shown in Fig. 10-107. This distribution permits the excitation of, or the removal of energy from the waveguide by introducing such an *aerial* or *probe* into the guide (Fig. 10-108). The probe is inserted at a position where the *E* lines are a maximum for removal of energy or where the *E* lines are desired to be a maximum for excitation. Coupling can be varied by moving the probe away from the position of maximum *E* lines or by exposing less of the length of the probe to the fields in the guide. Fig. 10-109 shows an arrangement for varying the amount of coupling. The position of the field with respect to the probe can be varied by moving plunger *A* in the end of the waveguide.

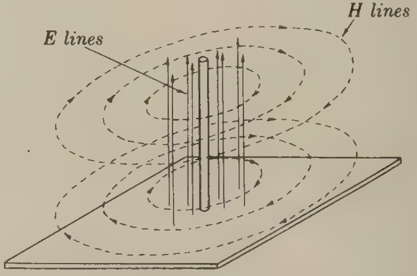


Fig. 10-107 *H* and *E* field distribution about a quarter-wave Marconi aerial.

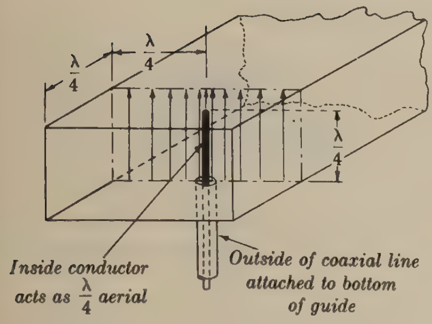

$\frac{\lambda}{4}$
$\frac{\lambda}{4}$
$\frac{\lambda}{4}$
Inside conductor acts as $\frac{\lambda}{4}$ aerial
Outside of coaxial line attached to bottom of guide

FIG.10-108 Probe coupling from a coaxial line to a waveguide.

Plunger *B* at the top of the extension of the coaxial line acts as a shorting bar to form a quarter-wave stub support for the centre conductor and also provides a means of tuning this section so that the electric field is at the proper position along the centre conductor to energize the waveguide. In effect, plunger *B* varies the effective length of the probe.

The probe method of energizing or extracting energy from a guide is usually superior to inductive coupling by means of a loop especially when large amounts of power are involved.

Both the loop method and probe method of coupling may be used to put energy into or to take energy from a waveguide. If the coupling device is adjusted for putting energy into a guide without producing reflections it is also well adjusted for taking energy out of the guide.

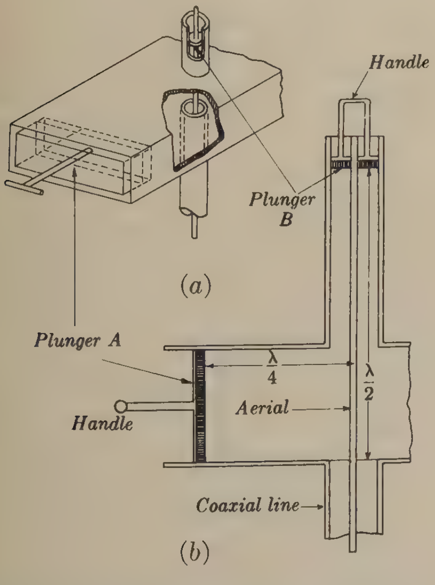


FIG.10-109 Adjustable probe coupling from a coaxial line to a waveguide (a) Perspective view (b) Cross-sectional view.

10.24 Characteristic impedance and matching. Each substance through which electromagnetic waves may pass offers a different resistance to the passage and may therefore be said to have a different characteristic impedance. For example, the impedance of free space is 377 ohms. This may be thought of as the voltage-current ratio of waves travelling in one direction. The effect of constraining the waves to pass through a waveguide affects this value. In general the impedance of a waveguide is greater than that of free space and that of a rectangular guide may be decreased by decreasing its *a* dimension. This latter is, however, impractical as the amount of power which it will then pass without arcing is very limited. Thus the impedance of a rectangular waveguide may vary from 0 to 456 ohms but the minimum characteristic impedance of a practical circular waveguide is of the order of 350 ohms.

Any abrupt change in the size or shape of a waveguide causes reflections as a result of the change in characteristic impedance. Thus to pass energy efficiently from a guide of one dimension to that of another, a flared matching section (Fig. 10-110) is used to permit the fields to expand gradually to fit the different sizes of guides.

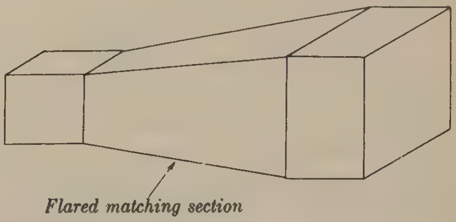


FIG.10-110 Flared matching section for a waveguide.

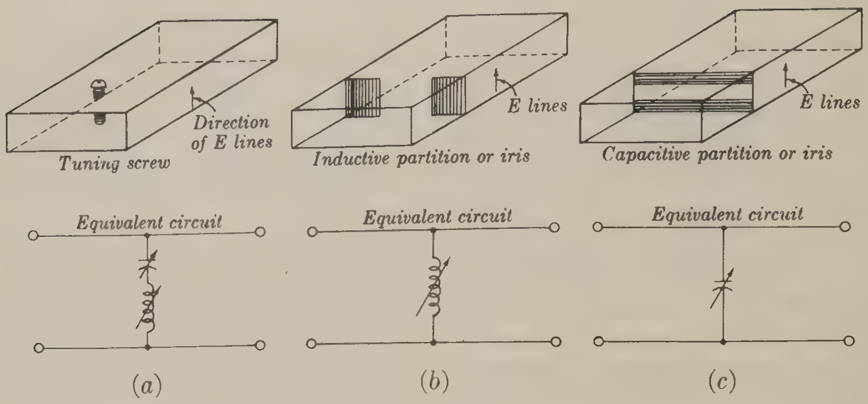


FIG. 10-111 Tuning screw and iris as matching devices in a waveguide and the equivalent circuits (a) Tuning screw (b) Inductive iris (c) Capacitive iris.

The characteristic impedance of a waveguide may also be changed by inserting a tuning screw in the top or bottom wall of a guide, Fig. 10-111 (a), or by inserting a perforated metal diaphragm called an *iris* across the guide as shown in Fig. 10-111 (b) and (c). Both the screw and the iris are equivalent to a shunt path across the waveguide. When either of these types of obstruction is completely removed from the waveguide, the reactance of the shunt path is infinite since the fields within the guide are not disturbed. If the opening in the iris is completely closed the electric field is zero at the surface of the metal blades and the shunt impedance is zero. This is not quite the case for the screw tuner as it does not present zero impedance even when it touches the bottom of the guide. The reactance of the shunt paths formed by the tuning screw or the iris depends upon the distance the device extends into the guide.

10.25 Waveguide joints.

A. Effect of Poor Joints

In practice a waveguide is usually fed directly from a magnetron oscillator. Thus the load on the magnetron is the impedance which the waveguide presents at its input end. If any irregularity occurs in the waveguide a reflected wave occurs and the effective impedance at the input end changes causing oscillation to take place at some undesirable frequency. This effect is called *frequency pulling*. In addition, if the waveguide is long an effect called *frequency splitting* may occur if the termination reflects a wave down the guide.

For these reasons it becomes necessary to design all joints or bends so that a minimum reflection occurs. Further, it is desirable to keep the total waveguide length as short as possible.

B. Mechanical Joints

At wavelengths of the order of 10 centimetres, coupling between two sections of waveguide is usually effected by means of flat end-flanges as shown in Fig. 10-112 (a) and (b). Fig. 10-112 (a) shows the two waveguide sections clamped together by means of end-flanges and bolts. The detail of each end-flange is shown in Fig. 10-112 (b). The principal cause of reflection at such joints as these results from misalignment of the walls of the guide at the joint. Such misalignment causes a step in the walls which results in reflections. To avoid this the bolt holes must be accurately located and care must be taken in making the connection. In addition, gaps between the waveguide walls across the joint

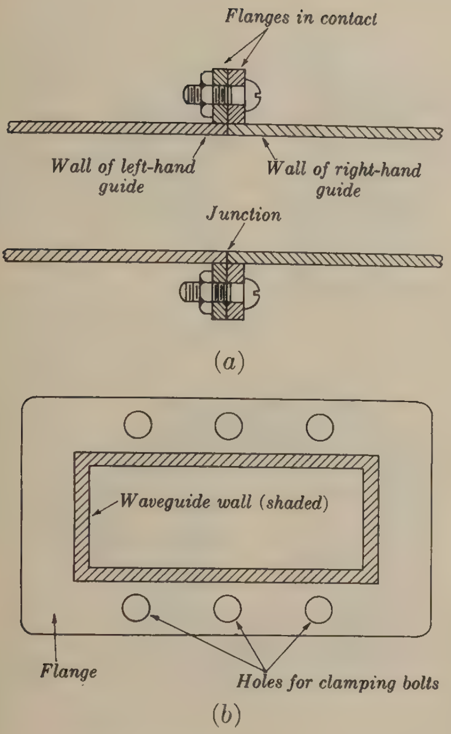


FIG. 10-112 Mechanical joint in a 10 cm. waveguide showing detail of end-flanges (a) Side view (b) End view.

may occur due to imperfections in the surfaces of the flanges. Fig. 10-113 shows an exaggerated view of this imperfection which results in reflections that are appreciable. Consequently the flange faces must be made accurately parallel, plane surfaces.

At wavelengths of the order of 3 centimetres these defects in coupling become even more difficult to avoid. Suitable coupling flanges are shown in Fig. 10-114 (a) and (b). The flanges are clamped together by means of a ring nut which presses on one and screws on the other. The walls of the guide sections are accurately lined up by means of accurately placed dowel pins.

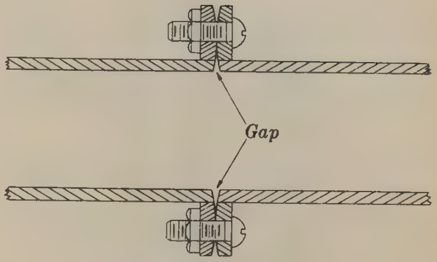


FIG. 10-113 Exaggerated view of gap between waveguide walls produced by irregularities in plane surfaces of end-flanges.

C. Choke Joints

As the frequency at which a waveguide is to operate is increased, the reflections caused by a gap between the walls of two jointed sections increase. However, a form of coupling which eliminates this defect makes use of the electrical properties of a gap between the flanges to provide a theoretically perfect electrical connection between two guide sections. Such a joint is called a *choke joint* and is shown in Fig. 10-115. This type of joint eliminates mechanical connection between the guide sections. As a result such a joint not only provides a good electrical connection but may also be used to isolate two sections of a system such as a waveguide and a radar transmitter so that vibration occurring in one part of the system does not occur in the other. The principle of operation of a choke joint may be understood by a consideration of Fig. 10-116. The joint consists of two flanges, one fixed to each waveguide section.

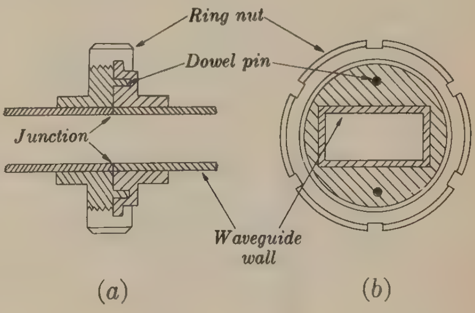


FIG. 10-114 Mechanical joint in a 3 cm. waveguide showing detail of end-flanges (a) Side view (b) End view.

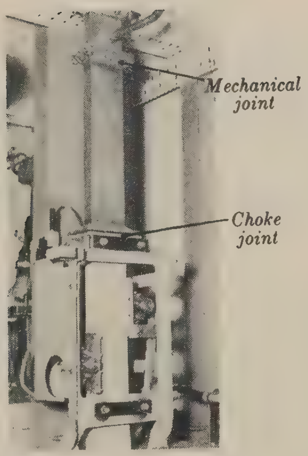


Fig.10-115 Choke joint and mechanical joint in a 10 cm. waveguide.

Mechanical isolation of the waveguide sections is accomplished by separating the flanges as shown. In practice this separation can be of the order of several millimetres without introducing appreciable reflection or loss.

The two flanges are parallel and their width is chosen so that they form a quarter-wavelength transmission line. When energy appears at the joint standing waves are set up in the quarter-wave section AB. The open circuit at A reflects a low impedance to the point B and a short circuit appears across CD. Thus, electrically, there is no break in the wall of the waveguide and the energy travels past the joint without suffering any disturbance.

Another type of choke joint is shown in Fig. 10-117 (a) and (b). In this case, the upper flange is a flat surface, while the lower flange has a slot cut in it one quarter wavelength deep. This slot is located so that the middle, M, of the long dimension of the waveguide is one quarter wavelength from the edge of the slot, Fig. 10-117 (b).

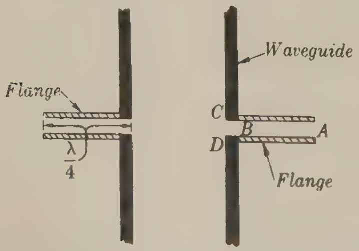


Fig.10-116 Diagram of a choke joint.

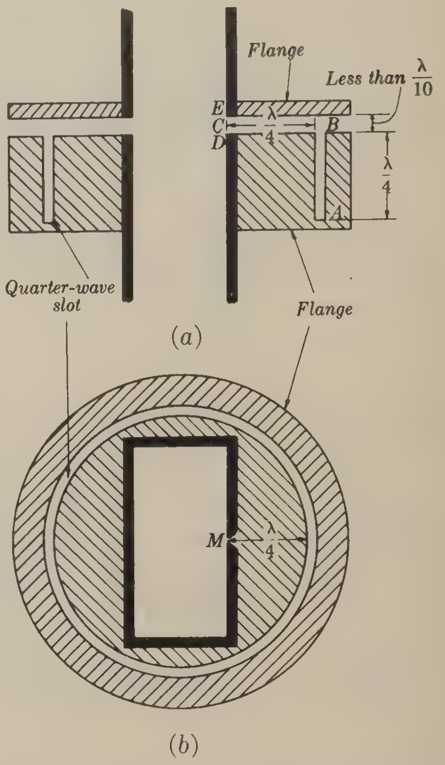


Fig.10-117 Choke joint connecting two sections of rectangular waveguide (a) Side view (b) End view.

When energy appears at the joint standing waves are set up in the quarter-wave sections AB and BC. The short circuit at A reflects a high impedance to B. This high impedance reflects a low impedance across the quarter-wave section from B to C. Thus a short circuit appears across DE. This same arrangement of a choke joint is also used with circular waveguides (Fig. 10-118). In addition to its other advantages such a joint may be used with circular waveguides to allow one section, such as

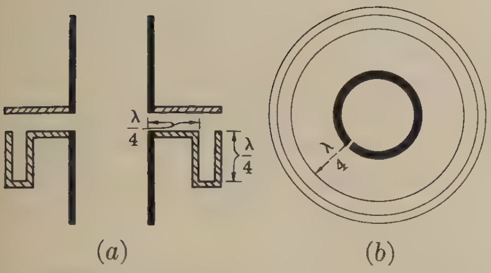


FIG.10-118 Choke joint in a circular waveguide (a) Side view (b) End view.

the aerial section, to turn relative to another section, the transmitter section. Thus a choke joint may be used in this way as a rotating joint for circular waveguides.

D. Rotating Joints

It has been seen that a choke joint may be used to connect two circular sections of waveguide. However, in practice, since the waveguide up to the rotating joint is usually rectangular and since the portion of the joint that rotates must be circular for mechanical reasons, a rotating joint involves a transition from the modes of a rectangular pipe to the modes of a circular pipe and vice versa. Such a system is shown in Fig. 10-119. Rectangular waveguide *A* is excited by the oscillator and operates in a $TE_{0,1}$ mode. A small exciting probe extends from a plug matching adjustment in this waveguide up into a vertical section of circular waveguide. This probe is parallel to the electric field in each waveguide as shown in Fig. 10-120.

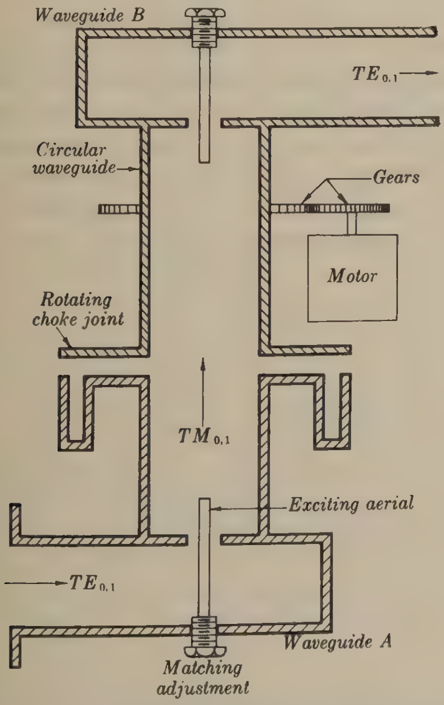


FIG.10-119 Rotating joint in a circular guide section connecting two rectangular guide sections.

Since the electric and magnetic fields in the circular waveguide operated in the $TM_{0,1}$ mode are symmetrical about the axis, rotation of one section of this guide relative to another section does not affect the overall polarization of the field. Thus a rotating joint inserted in the circular portion of the waveguide does not alter polarization. The rotating choke joint is similar to the one described previously in C. The upper rectangular waveguide, which rotates with the upper section of the circular waveguide, is excited in a $TE_{0,1}$ mode in a similar manner using probe coupling as in the

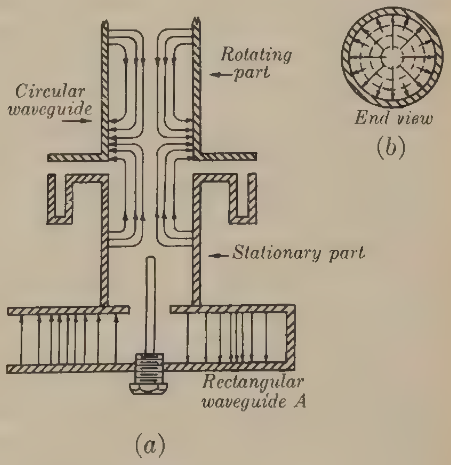


FIG.10-120 Electric field distribution in the rectangular and circular waveguides of Fig. 10-119 (a) Side view (b) End view of circular guide.

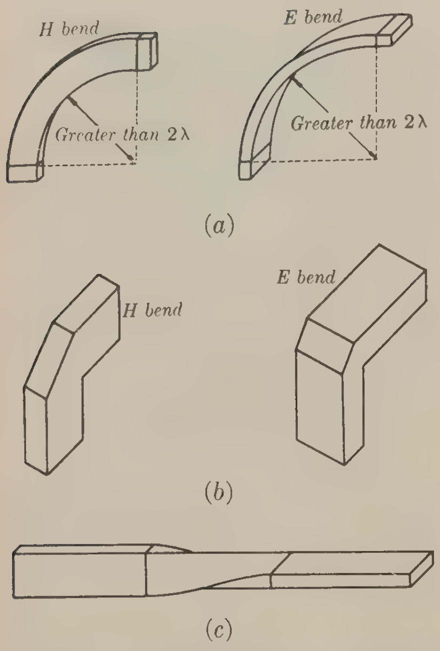


Fig.10-121 Bends and twists in waveguides (a) Curved H and E bends (b) Sharp H and E bends (c) 90-degree twist.

bottom waveguide. Matching adjustments are incorporated to change probe lengths.

10.26 Bends and twists. Often in practice it becomes necessary in a waveguide length or run to turn corners. This may be accomplished by bending the waveguide into a curve as shown in Fig. 10-121 (a). The waves travel around such an elbow without appreciable reflection provided the radius of curvature is of the order of 2λ. The bends shown are known as *H and E bends* respectively.

However, sometimes a sharp bend in the waveguide is imperative. Sharp elbows are shown in Fig. 10-121(b) in which reflection is reduced by beveling the elbow. The bevel is such that approximately one-half the square which would be formed if the edges of the waveguide were extended is removed. The bevel face inside the guide acts as a reflector.

Sometimes in connecting two differently oriented sections of waveguides it becomes necessary to rotate the field of one section so that there is proper matching to the second section. This can be done as shown in Fig. 10-121 (c). In this figure the field is rotated through an angle of 90°. Little reflection results provided the length of the twist is greater than 2λ.

10.27 Resonant waveguide sections. Waveguide sections which are closed at both ends or positioned in such a manner as to produce reflections, contain standing waves and act as resonant cavities. Such sections can be used for matching waveguides of different characteristic impedances as shown in Fig. 10-122.

Resonant sections may also be used as filters in a manner similar to that in which resonant sections of transmission lines are used for this purpose. Fig. 10-123 (a) shows a closed quarter-wave section of waveguide attached to a main waveguide. This section acts like a shorted quarter-wave stub and so presents a high impedance between its terminals. Thus in the equivalent circuit shown in Fig. 10-123 (b) there is a high impedance between A and B and the guide acts as if it had an open circuit between A and B for the frequencies at

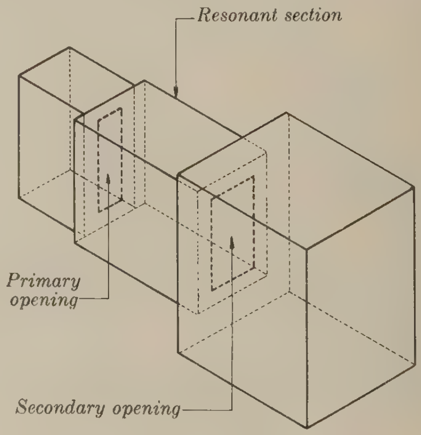


Fig.10-122 Resonant waveguide section as a matching device between two guides of different impedances.

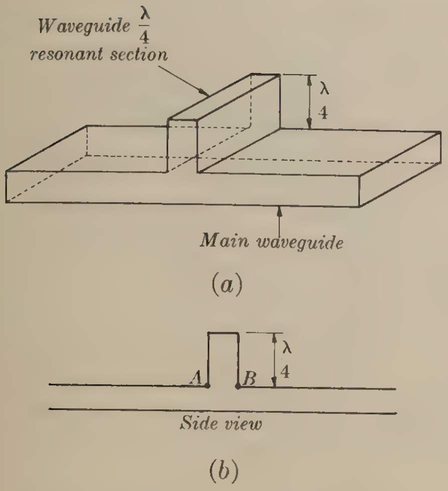


FIG. 10-123 Resonant quarter-wave guide section used as a wave filter (a) Perspective view (b) Side view.

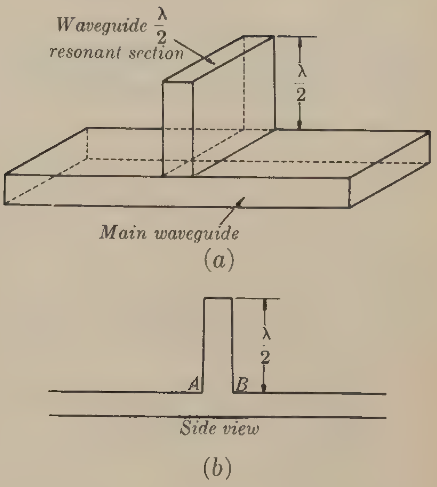


FIG. 10-124 Resonant half-wave guide section used as a wave filter (a) Perspective view (b) Side view.

which the stub resonates. For these frequencies the main waveguide presents a high impedance since the field distortion at the opening prevents their transmission.

Similarly, a half-wave closed section of waveguide may be attached to a main waveguide run as shown in Fig. 10-124. As for a half-wave shorted stub the resonant waveguide section presents a short across AB for the frequencies at which the guide resonates.

10.28 Energy radiation. If an energized waveguide is left open at the end it will radiate energy into space. However, the amount of energy radiated is small since the termination seldom provides a correct match to prevent standing waves and produce efficient transfer of energy. The waveguide may be flared into a horn as shown in Fig. 10-125 to match the impedance of the waveguide to that of free space and so produce a more efficient transfer of energy. Radiation from such horn devices is discussed in the chapter on aerials.

In addition to radiation from the end of an open waveguide, radiation also may take place from holes or slits in the guide. However, very little radiation takes place if such holes or slits are covered with wire screen.

10.29 A complete waveguide system. The component parts of a waveguide system are shown in Fig. 10-126. The various components of such a system as they occur in order from transmitter to aerial are indicated and labelled. Although the order of such components is not entirely rigid the order shown is typical of that used in many radar equipments.

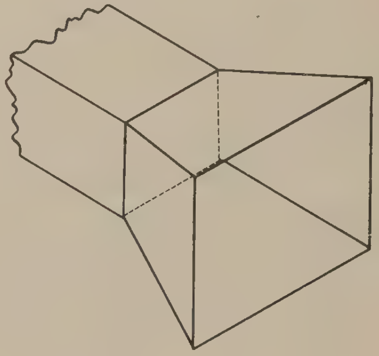


FIG. 10-125 Flared horn radiator used to match the waveguide impedance to that of free space.

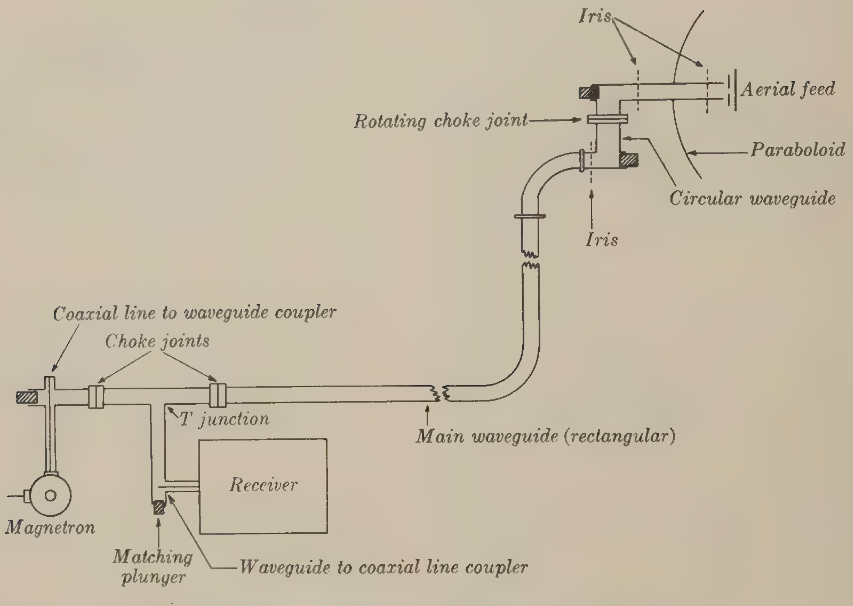


FIG. 10-126 Waveguide system in a radar set.

A. Rectangular Waveguides

λ_c (cm.) $TE_{0,1}$ MODE	INSIDE DIMENSIONS b(cm.)	INSIDE DIMENSIONS a(cm.)	OUTSIDE DIMENSIONS
4.57	2.29	1.02	½″ × 1″ 0.050″ wall
5.72	2.86	1.27	⅝″ × 1¼″ 1⁄16″ wall
14.44	7.22	3.40	1½″ × 3″ 0.080″ wall
14.00	7.00	0.95	15⁄32″ × 2⅞″ 1⁄16″ wall

B. Circular Waveguides

INSIDE DIMENSIONS d(in.)	INSIDE DIMENSIONS d(cm.)	λ_c (cm.) $TE_{1,1}$ MODE	λ_c (cm.) $TM_{0,1}$ MODE	λ_c (cm.) $TE_{2,1}$ MODE
15⁄16	2.38	4.06	3.11	
1 3⁄16	3.02	5.15	3.94	3.11
2½	6.35	10.83	8.29	
2¾	6.99	11.92	9.12	7.19
3	7.62	13.00	9.95	7.85
4	10.16	17.33	13.27	10.46

10.30 Standard sizes of waveguides. The standard sizes of waveguides are shown in the tables on page 10:68:

10.31 Summary of formulas for cut-off wavelengths. The formulas for determining the cut-off wavelengths for guides operating in the more common modes are given in the following table.

Formulas for Cut-Off Wavelength

RECTANGULAR WAVEGUIDES		CIRCULAR WAVEGUIDES	
MODE	CUT-OFF WAVELENGTH	MODE	CUT-OFF WAVELENGTH
$TE_{0,1}$	$2b$	$TE_{1,1}$	$1.71d$
$TE_{1,1}$ or $TM_{1,1}$	$\frac{2ab}{\sqrt{a^2+b^2}}$	$TM_{0,1}$	$1.31d$
$TE_{0,2}$	b	$TE_{0,1}$	$0.82d$
$TE_{1,0}$	$2a$	$TE_{2,1}$	$1.03d$
$TE_{m,n}$ or $TM_{m,n}$	$\frac{2}{\sqrt{\left(\frac{m}{a}\right)^2+\left(\frac{n}{b}\right)^2}}$	$TM_{1,1}$	$0.82d$

10.32 Questions and problems.

1. What is the purpose of a transmission line used
   (a) between a transmitter and an aerial
   (b) between a receiver and an aerial?
2. (a) Draw a diagram to illustrate the electric and magnetic fields about a simple vertical aerial at some instant when the a-c current in the aerial is a maximum.
   (b) Draw a similar diagram for an instant one period of the a-c later.
3. Using the diagrams of question 2 explain how an electromagnetic field is radiated away from the aerial.
4. What is the wave front of an e-m wave?
5. Using diagrams explain the meaning of the terms vertically polarized, horizontally polarized, obliquely polarized, circularly polarized, elliptically polarized, as applied to e-m waves.
6. What is the polarization of an e-m wave travelling over a highly conductive surface?
7. Why do standing waves of current and voltage exist on a long wire transmission line?
8. Draw a diagram to show the standing waves of voltage and current along an infinite two-wire line.
9. Why are the waves of question 8 called standing waves?
10. How far apart are successive voltage or current maxima in the diagram of question 8 if the a-c frequency is 50 Mc/s, 50 Kc/s, 50 c/s?
11. Why are the amplitudes of voltage and current reduced farther from the generator in the diagram of No. 8?
12. (a) What is a finite line?
    (b) What is an open-ended line?
    (c) What is a short-circuited line?

13. (a) What is the value of the current at the end of an open-ended line?
    (b) What is the value of this voltage?

14. How does reflection of the wave take place at the end of an open-ended line?

15. Explain how standing waves of current and voltage occur on an open-ended line.

16. How does the instantaneous value of voltage or current vary at a given point along an open-ended line?

17. How does the instantaneous value of voltage or current vary along the open-ended line at a given instant of time?

18. Explain the terms node, anti-node, loop.

19. How far apart along a line are nodes and loops?

20. Explain how the peak value of current or voltage varies at a given instant along a line for
    (a) a travelling wave
    (b) a standing wave.

21. (a) What is the standing-wave ratio of a line?
    (b) The standing-wave ratio of a line is 2:1. If the voltage measured at a voltage anti-node is 18, what is the voltage at a node?

22. Voltage nodes occur along a two-wire parallel conductor line at intervals of 0.425 feet. Calculate the frequency of the wave in Mc/s.

23. Along a concentric line with air insulation the voltage anti-nodes are 12.7 cm. apart. Calculate the frequency of the wave in Mc/s.

24. The frequency of a generator feeding a twisted-pair line is 2000 Mc/s. Calculate the possible minimum and maximum values of the distance between successive current nodes along the line.

25. (a) What is an absorption-type standing-wave indicator?
    (b) What are the components of a simple indicator of this type?
    (c) How must an absorption-type indicator be used to indicate (i) current nodes and loops (ii) voltage nodes and loops?

26. Explain the action of the indicator shown in Fig. 10-12.

27. (a) Explain the action of the indicator of Fig. 10-13.
    (b) What are the functions of R_1 and R_2, L_3 and L_4?

28. What are the values of current and voltage at the end of a shorted line?

29. What is the requirement for the production of standing waves along a line?

30. (a) What are the four electrical constants of a two-wire line?
    (b) Draw a diagram to illustrate how a two-wire line is considered to consist of these constants.

31. (a) Explain what is meant by the characteristic impedance of an infinite line.
    (b) What symbol represents it?

32. What is the input impedance of a short length of line with an output load equal to the characteristic impedance?

33. Explain what is meant by the characteristic impedance of a line of definite length.

34. What is the condition for maximum transfer of energy from
    (a) a line to a load
    (b) a generator to a line?

35. Why are standing waves not present along a line terminated by its characteristic impedance?

36. Give the formula for the characteristic impedance of a two-wire line, explaining the meaning of each symbol.

37. Draw a diagram to show the distribution of current and voltage on a dipole.

38. Explain why the dipole radiates efficiently while a two-wire line radiates very little.

39. How does the impedance vary along a dipole?

40. (a) What is the character of the impedance at the centre of a dipole?
    (b) What is its value in the case of a dipole situated far from disturbing influences?
    (c) What is it called?

41. (a) What is a non-resonant line?
    (b) What should the value of the standing-wave ratio on such a line be?
    (c) Compare the phases of current and voltage on this type of line.

42. Explain the advantages of a non-resonant line when used
    (a) between a transmitter and aerial
    (b) between an aerial and a receiver.

43. How is a transmission line made non-resonant?

44. (a) Name two kinds of match to a half-wave aerial.
    (b) How is a match indicated?

45. Explain the delta match to a half-wave aerial using a diagram to illustrate.

46. Draw a diagram and explain the operation of
    (a) a quarter-wave stub match to an aerial
    (b) a half-wave stub match to an aerial.

47. Why are there two possible matching points when a half-wave stub match to an aerial is used?

48. (a) What is a resonant line?
    (b) Explain its advantages and disadvantages.

49. Compare a two-wire resonant line and a two-wire non-resonant line with respect to matching, heat losses, radiation losses and need for insulation.

50. Name two types of resonant-line feed to an aerial.

51. Using diagrams explain the operation of a resonant-line centre feed to an aerial with the transmitter connected at
    (a) a current node
    (b) a current loop.

52. Using a diagram explain the operation of a resonant-line end feed or an aerial with the transmitter connected at
    (a) a current node
    (b) a current loop.

53. Explain how a resonant line is used to feed a half-wave aerial which is to be used at the first, second or third harmonic frequencies.

54. Under what conditions does a resonant-line section act like a conventional resonant circuit?

55. Draw diagrams to illustrate the current, voltage and impedance distributions along a 2λ resonant line section when the line is
    (a) open-ended
    (b) short-circuited.

56. Using the diagrams of Question 55, fill in the table at the bottom of this page for an open-ended resonant line.

57. Make a table similar to that of question 56 for a short-circuited resonant line.

58. Draw diagrams to show the change in reactance along a 2λ resonant-line section when the line is
    (a) open-ended
    (b) short-circuited.

59. Make tables to show the type of reactance for the following lengths of resonant-line sections: λ/8, less than λ/4, $\frac{3}{8}\lambda$, $\frac{5}{8}\lambda$, $\frac{7}{8}\lambda$, when the line is
    (a) open-ended
    (b) short-circuited.

60. Give the chief characteristic of
    (a) a quarter-wave section
    (b) a half-wave section.

61. (a) Name six types of transmission lines.
    (b) Briefly describe the construction of each.

62. (a) What are the advantages and disadvantages of each type of line named in question 61?
    (b) Discuss the applications of each type.

63. Give the formula for the characteristic impedance of the following lines, explaining the significance of each symbol:
    (a) single-wire line
    (b) two-wire line with air dielectric
    (c) four-wire line with air dielectric
    (d) rigid air-dielectric concentric line
    (e) flexible coaxial line with any dielectric.

64. Calculate the characteristic impedance of a single-wire line 0.5 cm. in diameter and 30 feet from the ground.

65. (a) Give the formula for the characteristic impedance of any line in terms of its inductance and capacitance.
    (b) Give formulas for the inductance and capacitance of a two-wire line.
    (c) Using the formulas of (a) and (b) find a formula for the characteristic impedance of a two-wire line.

66. Calculate the characteristic impedance of a two-wire line made of copper tubing $\frac{3}{8}$ inch in diameter spaced 4 inches apart. Check the result by using the chart of Fig. 10-37.

67. Use the chart of Fig. 10-37 to find the necessary spacing of a two-wire line of No. 16 wire to give a characteristic impedance of 500 ohms.

68. Check the value of the answer obtained in question 67 by using the formula.

LENGTH OF LINE SECTION	IMPEDANCE	TYPE OF EQUIVALENT RESONANT CIRCUIT
λ/4		
λ/2		
$\frac{3}{4}\lambda$		
1λ		
$1\frac{1}{4}\lambda$		
$1\frac{1}{2}\lambda$		
$1\frac{3}{4}\lambda$		
2λ		

69. (a) Use the chart of Fig. 10-47 to find the characteristic impedance of a rigid air-dielectric concentric line in which the inner diameter of the outer conductor is 0.75 inches and the outer diameter of the inner conductor is 0.25 inches.
    (b) Check this value by using the formula.

70. What is the effect on the characteristic impedance of a flexible coaxial line if the dielectric is a solid material other than air?

71. (a) Explain the purpose of the components of the transmission-line oscillator of Figs. 10-48 and 10-49.
    (b) At what frequency is it designed to operate?
    (c) What kind of resonant-line section is used and what is its function in the oscillator circuit?
    (d) What conditions of construction are necessary for the *Q* of a two-wire line or a concentric line to have an optimum value for use of the line in a transmission-line oscillator?

72. (a) What are Lecher lines?
    (b) What are the advantages of Lecher lines used as oscillatory circuits in oscillators?

73. Give seven possible applications of resonant line sections.

74. (a) Explain how a quarter-wave shorted section may be used as an insulating device.
    (b) Describe specific cases where this principle is employed.

75. Explain how a quarter-wave section is used as a wave filter to pass
    (a) even harmonics
    (b) odd harmonics of the fundamental frequency.

76. Use a diagram to explain how a quarter-wave shorted section is used to match
    (a) a resonant line to a non-resonant line
    (b) a receiver to an input line.

77. An aerial is to be matched to a non-resonant line. Describe how this is accomplished using a half-wave shorted resonant line.

78. (a) What is a quarter-wave transformer?
    (b) Explain how it is used to match a 650Ω line to a 450Ω line.

79. What spacing is required for a two-wire air-dielectric quarter-wave transformer made of copper wire 0.1 inches in diameter which is to be used to match a 600Ω line to a 72Ω dipole?

80. Explain how a half-wave open-ended section may be used to match a 550Ω line to a 650Ω line.

81. (a) Using a diagram describe the operation of a single-stub tuner used to match a line to an aerial.
    (b) Explain how the position of the stub connection is found.

82. Explain how a stub match is used to match a coaxial line to an aerial.

83. Two aerials are to operate 180° out of phase. Explain how a non-resonant line is used to feed these aerials.

84. Explain the operation of the feeding arrangements shown in Fig. 10-69 and Fig. 10-70.

85. (a) What kind of resonant-line section is used as a phase inverter?
    (b) Explain how this is possible.

86. (a) Why are coaxial-line balance converters required?
    (b) Name two types of line converters.

87. With the aid of a diagram describe how a sleeve transformer is used to convert an unbalanced coaxial line to a balanced line.

88. Explain the operation of the trombone transformer when used to transform an unbalanced coaxial cable to feed a balanced dipole. Draw a diagram to aid the explanation.

89. What is a rotating joint and when is it necessary?

90. How is inductive coupling used with a rotating joint?

91. (a) Explain the operation of the rotating-joint couplings illustrated in Fig. 10-77 and Fig. 10-78.
    (b) How is the quarter-wave section made use of in each case?

92. What are
    (a) guided waves
    (b) unguided waves?

93. Why does almost all the energy in a guided wave reach its destination?

94. In microwave phenomena, how is the transmission of energy along a transmission line regarded as taking place?

95. In a sinusoidal plane-polarized electromagnetic wave travelling in free space
    (a) what fields are present
    (b) how are the fields situated?

96. What is a type *TEM* wave?

97. What is the velocity of propagation in a dielectric medium?

98. What are the characteristics of a dominant mode?

99. Which type of transmission line is usually referred to as a waveguide?

100. Compare the losses in a hollow waveguide with those in an open two-wire line and coaxial line.

101. What is the formula for power transmitted on a coaxial cable?

102. Why is the waveguide capable of transmitting higher power than the coaxial cable?

103. What advantages are there in waveguides as compared to other transmission lines with respect to construction?

104. How can a waveguide be considered to be formed by quarter-wave sections?

105. What frequencies may a waveguide carry?

106. Explain the range of frequencies possible for the waveguide when considering Figs. 10-87 and 10-88.

107. What is meant by critical minimum frequency or cut-off frequency?

108. Give the relative importance of dimensions *a* and *b* in Fig. 10-88.

109. What are practical dimensions for a rectangular waveguide for 10 cm. application?

110. Describe the representation of electric lines in Fig. 10-90.

111. Describe the arrangement of the *E* lines in Fig. 10-91.

112. What are the arrangements of the *H* lines in Figs. 10-93, 10-94, 10-95?

113. What is the *H* line distribution in Fig. 10-96?

114. What general methods are employed to show *E* and *H* lines clearly in a diagram of fields in a waveguide?

115. What is meant by a mode of operation?

116. Describe the system of numbering to differentiate between the modes of operation.

117. What type of mode is referred to as a magnetic wave?

118. How is the system of numbering applied to a circular waveguide?

119. Explain why the modes indicated in Fig. 10-99 are determined correctly from the figure.

120. Describe how a waveguide with a $TE_{0,2}$ mode is developed from a two-wire transmission line in Fig. 10-100.
121. How are circular waveguides developed from two-wire transmission lines in Fig. 10-102?
122. Give the formula for the maximum wavelength which can be transmitted in the circular waveguide.
123. How does the velocity of electromagnetic waves in a guide compare with the velocity in free space?
124. What is meant by group velocity?
125. What is meant by phase velocity?
126. What are the conditions of phase and group velocity when the r-f is at the cut-off frequency?
127. What are the two principal methods by which energy is put into, or removed from, a waveguide?
128. Describe the three methods of coupling from a coaxial cable to a waveguide in Fig. 10-106.
129. How is a probe employed with a waveguide for maximum excitation or removal of energy?
130. Describe the arrangement for varying the amount of coupling in Fig. 10-109.
131. What is the characteristic impedance of free space?
132. How does the impedance of a waveguide compare with that of free space?
133. How may the impedance of a rectangular waveguide be decreased?
134. What is the minimum characteristic impedance of a practical circular waveguide?
135. What effect does change in shape or size of a waveguide have on its characteristic impedance?
136. How is energy passed efficiently from a guide of one size to that of another?
137. Give methods for changing the characteristic impedance of a waveguide.
138. Describe what is meant by frequency pulling and how it may be caused.
139. What is frequency splitting and what is its cause?
140. Describe how mechanical joins between waveguide sections are achieved with minimum reflection.
141. What is meant by a choke joint?
142. Why is a choke joint employed?
143. Describe the principle of operation of the choke joint in Fig. 10-116.
144. Give the construction and operation of the choke joint in Fig. 10-117.
145. What special advantage is there in a choke joint with circular waveguides?
146. Describe the arrangement and operation of the rotating joint shown in Fig. 10-119.
147. What are the waveguide arrangements employed for turning corners?
148. How are waveguide sections employed as resonant cavities?
149. For what purposes may resonant cavities be employed?
150. What is the result if an energized waveguide is left open at the end?
151. How may the open end of the waveguide be adjusted to produce an efficient transfer of energy into space?
152. How may energy be radiated from the side of the guide?
153. Describe the arrangement and purpose of the parts of the waveguide system shown in Fig. 10-126.
154. (a) Give the formula for the cut-off wavelength for a $TE_{m,n}$ mode.
    (b) How is the above formula modified for the modes, $TE_{1,0}$, $TE_{0,2}$, $TE_{1,1}$, $TE_{0,1}$?

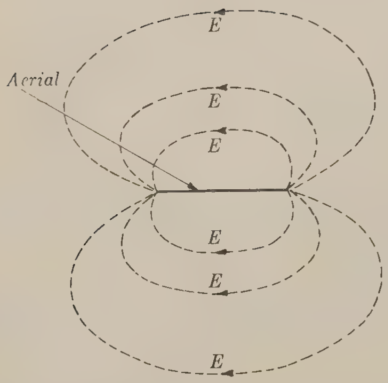


Fig.11-2 Electric field waves produced by the standing waves of current.

pended either horizontally or vertically depending upon the type of polarization and directivity desired.

A. The Horizontal Dipole

Consider a half-wave aerial suspended horizontally above the ground. Fig. 11-3 illustrates the aerial as viewed from above. Suppose an observer moves around a circle, the centre of which is the middle of the dipole, and measures the field strength. At a position at right angles to the line of the dipole the field strength is observed to be a maximum. The field strength decreases and becomes zero at a position opposite to the end of the dipole. As the observer continues to move, the field strength increases and becomes a maximum again at right angles to the line of the dipole. Therefore, if the observer changes position and moves so as to obtain constant field strength, he would move nearer the aerial as he approaches the line of the dipole, and away from the aerial as he approaches a direction perpendicular to the line of the dipole. A path of constant field strength is a circle on one side of the dipole. A similar path is obtained on the other side of the dipole. Thus, the radiation pattern for the horizontal dipole is a figure of eight as in Fig. 11-3.

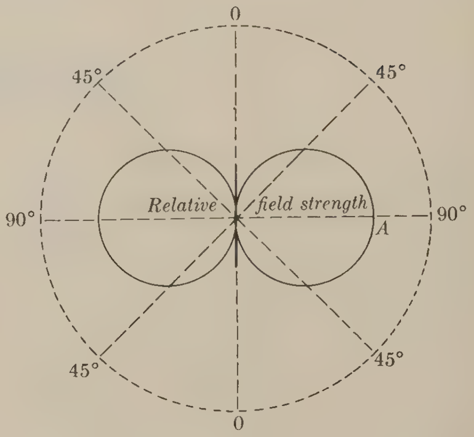


Fig.11-3 Radiation pattern of horizontal dipole.

The same radiation pattern is obtained in the form of a diagram in the following manner. The observer moves in a circle the centre of which is the centre of the dipole. At given positions around the circle, expressed in degrees relative to the line of the dipole, values of field strength are noted. For each angular position, a point is plotted on a diagram. The distance of the point from the centre of the diagram is made proportional to the field strength. The points are joined to form a diagram as in Fig. 11-3. This is called a *polar diagram.*

In practice this diagram is not exactly obtained if the aerial is near ground and other objects. The diagram in Fig. 11-3 is obtained theoretically, assuming the aerial to be remote from ground and other objects, that is, in free space.

If the radiation in all directions from the horizontal to the vertical is considered, a three-dimensional radiation pattern is obtained as in Fig. 11-4. Thus, a complete surface pattern of the radiation from the dipole resembles a doughnut. Fig. 11-5 shows a cross-sectional view of this pattern.

The simple horizontal free-space directive diagram (Fig. 11-2) of an aerial does not give the complete picture of the performance of the aerial since the waves

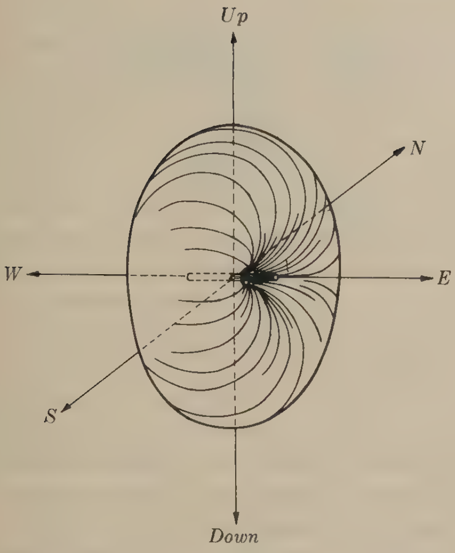


FIG.11-4 Complete surface radiation pattern of horizontal dipole.

usually of value for communication depart at some angle above the horizontal. Thus there may be useful radiation off the ends of the horizontal dipole at some angle above the horizontal. For example, in Fig. 11-5, *OA*, *OB* and *OC* all point in the same geographical direction, east, but each is at a different angle to the horizontal. The purely horizontal wave along *OA* has zero intensity, but *OB* and *OC* are of appreciable intensity. The higher the angle considered the greater the field intensity. Thus in plotting such a polar diagram it is necessary to specify the angle of radiation for which the diagram applies. The shape of the diagram is different for different radiation angles. Fig. 11-6 shows the polar diagram of a horizontal dipole at radiation angles of 10, 15 and 30 degrees.

For the horizontal dipole, the following facts are important.

(i) Directional effects occur in the horizontal plane.
(ii) No directivity occurs in the vertical plane at right angles to the dipole.
(iii) Radiation is horizontally polarized.

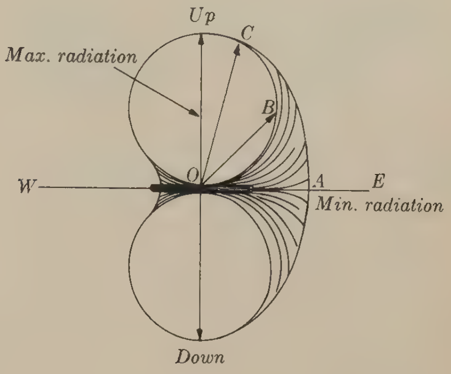


FIG.11-5 Cross-sectional view of surface radiation pattern of horizontal dipole.

The following terms associated with the study of directional aerials are important.

1. GAIN.

The merit of a directional aerial is measured in terms of its gain. The gain is defined as the ratio of the power that must be supplied to a standard comparison aerial to lay down a given signal at a given point, to the power that must be supplied to the directional aerial to give the same signal strength at the same point.

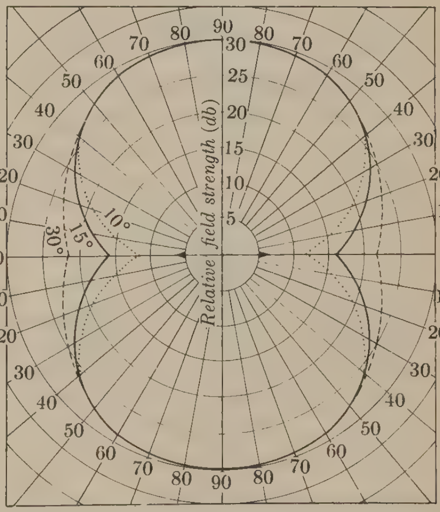


FIG.11-6 Polar diagram of horizontal dipole at radiation angles of 10, 15 and 30 degrees.

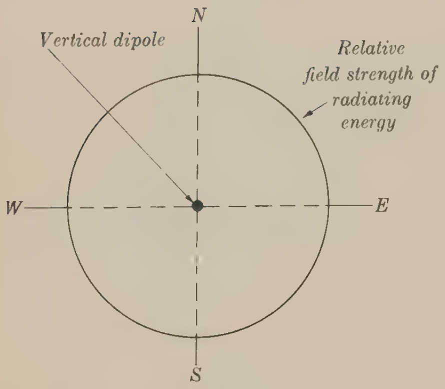


Fig.11-7 Polar diagram of radiation from a vertical dipole.

2. Directivity.

The directivity of an aerial refers to the sharpness or narrowness of the radiation pattern. The sharper the pattern the greater the directivity. Directivity usually results in an increase in gain but some aerials are capable of producing increased directivity with little or no increase in gain. Aerials with sharp patterns in the horizontal plane are said to have good *horizontal directivity*, while aerials with sharp patterns in the vertical plane are said to have good *vertical directivity*.

B. The Vertical Dipole

Consider the half-wave aerial suspended vertically in free space (Fig. 11-7) and again imagine that the aerial is being looked down upon from a height with the page representing the ground. The variation of signal strength around the aerial is represented graphically by the polar diagram shown in Fig. 11-7. This directive pattern in the horizontal plane is simply a circle indicating that the radiation is the same in all directions. A complete surface pattern of the radiation from this aerial resembles a doughnut as shown in Fig. 11-8 and Fig. 11-9. For any given angle in the vertical plane the field strength is the same in any compass direction. The field strength is different for each vertical angle of radiation but the shape of the polar diagram does not change when the angle is changed. Thus there is no horizontal directivity. Theoretically the vertical dipole in free space has no vertical radiation. However, it may produce considerable radiation at small angles to the vertical and so produce considerable upward radiation. For example, in Fig. 11-9, radiation along OA is zero but along OB and OC the radiation is appreciable. Because of this variation in field strength at different vertical angles, a polar diagram of the vertical aerial must specify the vertical angle of radiation for which the pattern applies.

For the vertical dipole, the following facts are important.

(i) Directional effects occur in the vertical plane.

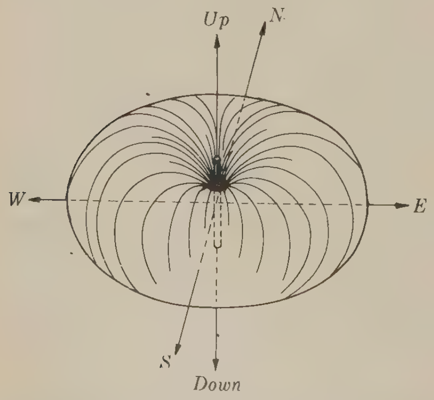


Fig.11-8 Complete surface radiation pattern of vertical dipole.

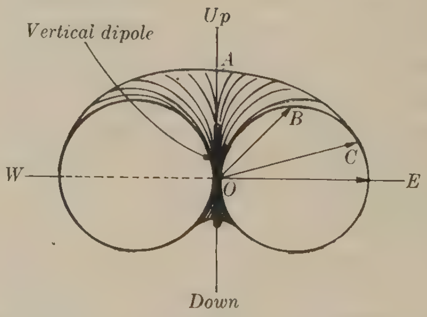


Fig.11-9 Cross-sectional view of surface radiation pattern of vertical dipole.

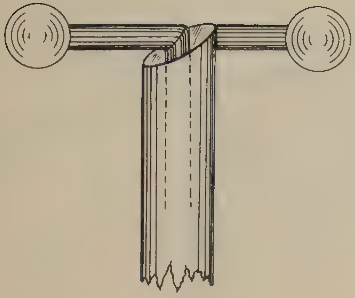


FIG.11-10 Special shapes of dipoles.

(ii) No directivity occurs in the horizontal plane.

(iii) Radiation is vertically polarized.

Dipoles may have special shapes as shown in Fig. 11-10. Such large diameter aerials are insensitive to small changes in frequency and are not so subject to corona discharge as are cylindrical dipoles.

11.2 Modified dipoles.

A. The Folded Dipole

The single dipole presents an impedance of approximately 73Ω at its centre (Sec. 10.7). This low-impedance input is preferable when a low-impedance transmission line is used to feed the dipole. However, when an open-wire transmission line is to be used a dipole having a higher input impedance is desirable. This is accomplished by using a folded dipole (Fig. 11-11 and Fig. 11-12) which consists of two parallel closely-spaced dipoles connected together at the ends. The folded dipole behaves as two parallel dipoles carrying equal currents. This can be understood by observing in Fig. 11-12 that the voltage at the ends of the two dipoles is held the same by the end connections and therefore the distribution of voltage along the dipoles is the same. Since the dipoles are similar, equal currents accompany the equal voltages. Thus the total length of the radiating element is now approximately one wavelength and the aerial is fed at the ends, a point of high voltage and low current.

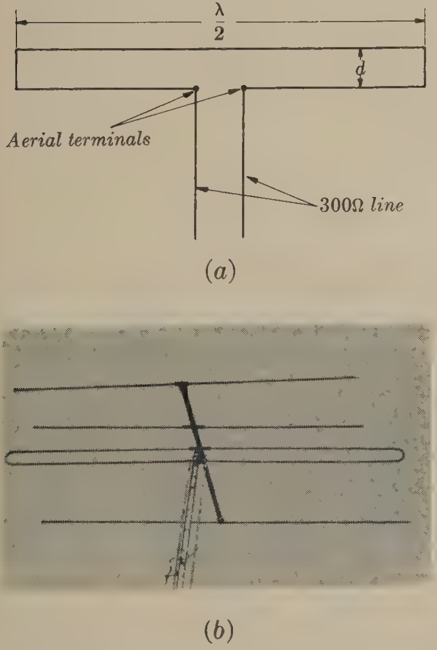


FIG.11-11 The folded dipole.

If, in a simple dipole and a folded dipole the currents are both I, then the following comparison is made.

For simple dipole

Current $= I$ Field $= E$

Input power $= P$

Input impedance $= \frac{P}{I} = Z$

For folded dipole

Current $= I$ Field $= 2E$

Input power $= 4P$

Input impedance $= \frac{4P}{I} = 4Z$

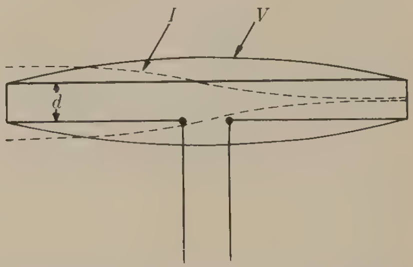


FIG.11-12 Voltage and current distribution on folded dipole.

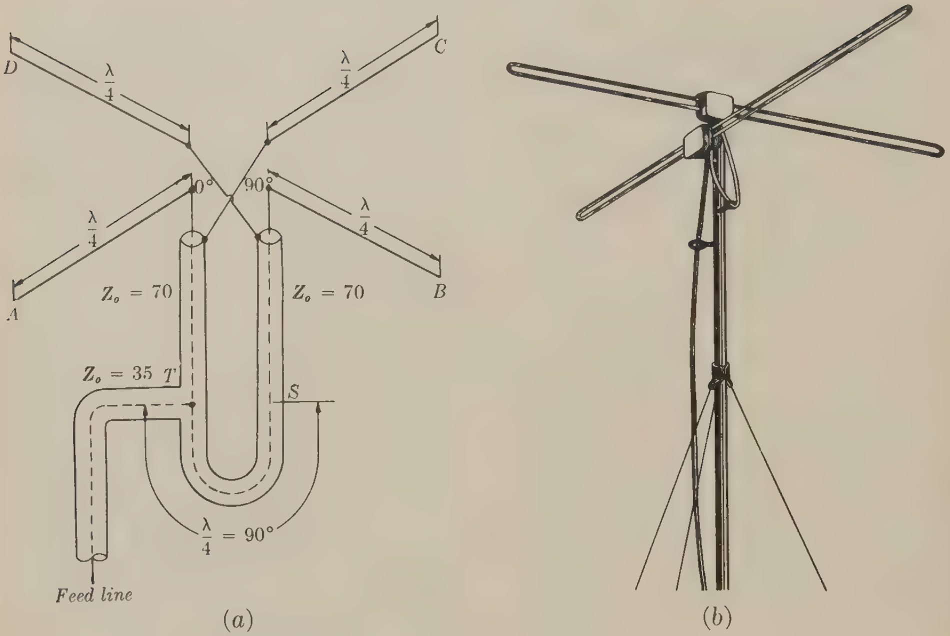


FIG. 11-13 The turnstile aerial.

A folded dipole has an input impedance four times that of an ordinary dipole. Therefore with the aerial in free space the input impedance is approximately 300Ω. It is practical to connect an open-wire transmission line of approximately 300Ω directly to the terminals of the aerial without any matching transformer.

The spacing, d, between the rods is of the order of $\frac{\lambda}{100}$ for wavelengths of approximately one-half meter.

The directive pattern of a folded dipole is similar to that for a single dipole.

B. The Turnstile Aerial

The turnstile aerial, Fig. 11-13 (a) and (b), consists of two driven dipoles erected horizontally at right angles to each other. The dipoles are fed in such a way as to give a 90° phase shift between the dipoles. This is accomplished by means of a trombone transformer. Element AC is across one concentric line and element BD is across the other. The characteristic impedance of each of these lines is matched to the input impedance of each of the dipoles to eliminate standing waves on the lines. In effect the two branching transmission lines are fed in parallel at the tapped point T. The transmission line feeding this point has one-half the characteristic impedance of the other lines to prevent standing waves on the fed line.

Since the distance between points T and S is $\frac{\lambda}{4}$, the energy at S lags that at T by 90°. The distance from S to dipole BD is exactly the same as the distance from T to dipole AC. Hence radiation from dipole BD lags that from AC by 90°. The field strength broadside to one of the dipoles is equal to the field from that dipole alone (Fig. 11-14). The field strength at any other angle is equal to the vector sum of the fields from the two dipoles at that angle. The vector sum of the fields from both dipoles at any angle is nearly equal to the field strength when measured broadside to one dipole. Hence a nearly circular horizon-

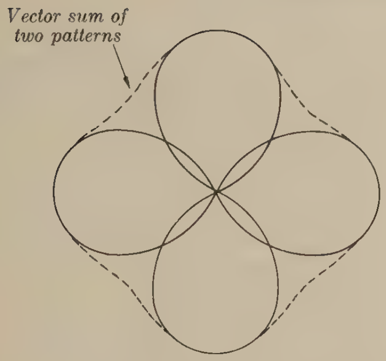


FIG. 11-14 Horizontal directive pattern of turnstile aerial.

tal directive pattern is produced as shown.

C. The Circular Dipole Aerial

The *circular dipole* aerial (Fig. 11-15) consists of three dipoles bent to form a circle. The three dipoles are driven in phase and the radiation is approximately uniform in all directions. The interaction of the fields is slightly greater than for the turnstile aerial. The interaction is uniform and does not detract from the usefulness of the system as a non-directional aerial. The dipoles are centre fed and a sleeve transformer is used to enable the unbalanced coaxial line to feed the balanced dipoles.

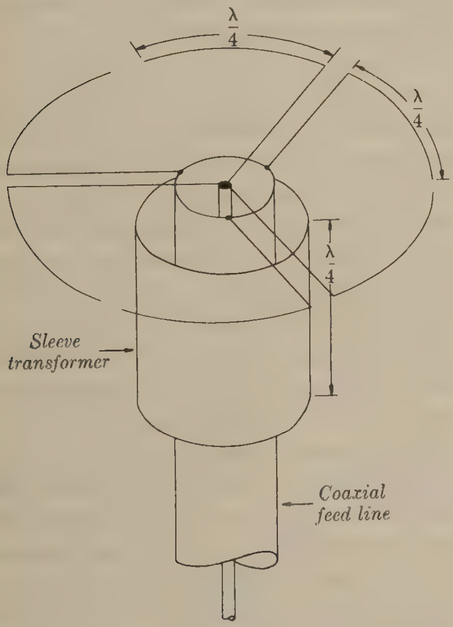


FIG. 11-15 The circular dipole.

D. Conical Cage Type Dipole

This aerial (Fig. 11-16) provides an all round radiation and uniform efficiency over a band of frequencies, for example, the band 155-185 Mc/s.

The aerial is a vertical dipole, each half

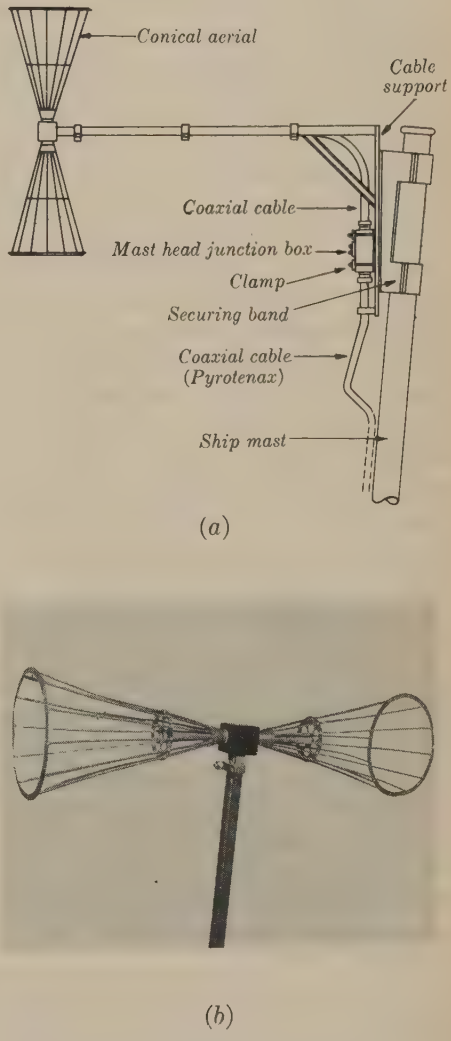


FIG. 11-16 Conical cage type dipole.

of which is cone shaped. To obtain best results it is fitted as high on the mast as possible and well isolated from all metal work. To minimize the shielding effect of the mast the aerial is connected to an L-shaped length of steel-sheathed twin-feeder cable so that it can be clamped to a gibbet-shaped structure which, when fastened to the mast, supports the dipole with its axis vertical in a position about four feet away from the mast. In order to obtain uniform efficiency over a band of frequencies it is important that the aerial load impedance remain constant within reasonable limits. The larger the surface area of the two elements of the dipole the less critical is the distribution of current and voltage with frequency and therefore the less the variation of the input impedance. By constructing the two elements conical in shape a large surface area is obtained and minimum variation in efficiency over a band of frequencies results. The main feeder is usually matched to the aerial at the upper end of the frequency band and is only slightly mis-matched at the lower end of the band.

11.3 Phased systems. The three main types of phased systems are:

(i) the collinear aerial
(ii) the broadside aerial
(iii) the end-fire aerial.

A. The Collinear Aerial

The *collinear aerial* consists of two or more half-wavelength elements placed end to end and excited in phase. The simplest form of collinear aerial consists of two horizontal half-wave elements erected so that they are in line. To understand the directivity of such an aerial consider first the standing waves of current which exist on an aerial consisting of three half-wavelength elements as shown in Fig. 11-17 (a). The current is opposite in phase in each half wavelength. Now if the aerial is broken at its centre and quarter-wavelength sections are turned back as shown in Fig. 11-17 (b) currents in the half-wavelength sections are in phase. Thus the waves reinforce along the horizontal plane at right angles to the aerial and the directive pattern is as shown in Fig. 11-17 (c). Radiation from the stub is negligible.

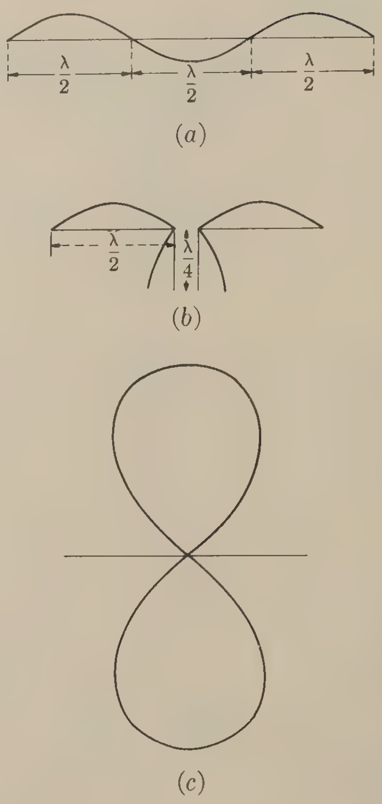


Fig.11-17 Standing waves of current on a $\frac{3\lambda}{2}$ aerial and directive pattern.

The collinear aerial has a directive pattern similar to that of the simple dipole but with added directivity. Increased directivity is obtained from the three-element aerial, Fig. 11-18 (b), and from the four-element aerial, Fig. 11-18 (c). These aerials require the use of quarter-wave phasing sections as shown. The directive patterns for all three aerials are shown in Fig. 11-18 and it should be noted that the beam becomes

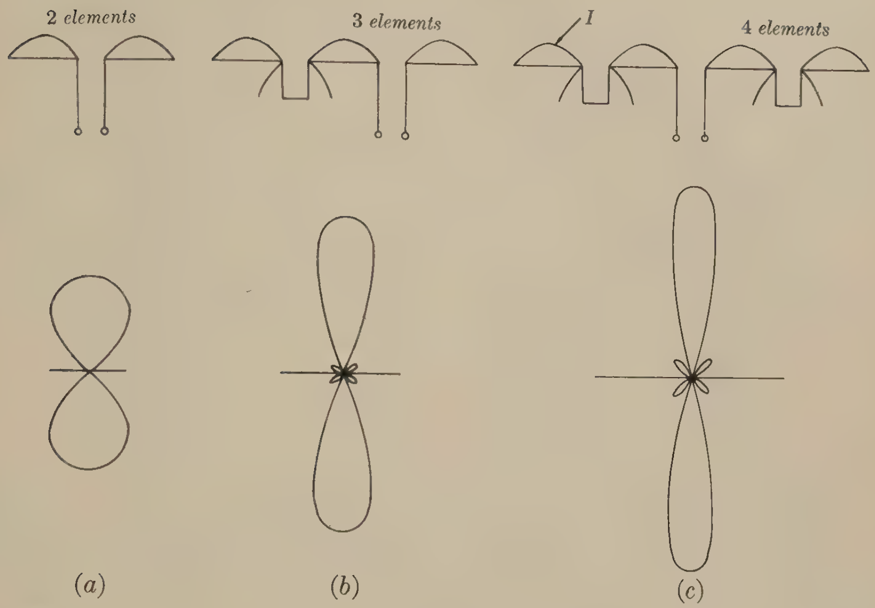


FIG.11-18 Directive patterns of collinear aerials.

sharper as the number of elements is increased.

Two half-wavelength elements give a power gain approximately 1.6 times greater than that of the simple dipole. The three-element aerial has a power gain of approximately 2.6, the four-element array approximately 3.2.

When the collinear aerial is erected horizontally it is directive in the horizontal plane; when erected vertically it is vertically directive. All collinear aerials are bi-directional.

B. The Broadside Aerial

The simple *broadside aerial* consists of two half-wave radiating elements mounted parallel to each other a half wavelength apart and driven so that their currents are in phase. The principle of the two-element broadside aerial may be understood by a consideration of Fig. 11-19. Two half-wave elements *A* and *B* are erected vertically one half wavelength apart. Imagine that these elements are being looked upon from the side with the page representing the plane of the aerials. Since the currents in *A* and *B* are equal and in phase the fields radiated from the elements at any instant are the same in polarity, phase and amplitude. By the time energy radiated from *A* reaches *B*, because of the time required for the energy to travel the half wavelength from *A* to *B*, the phase of this energy in the field at *B* is 180° out of phase with the energy radiated from *B* and the two fields cancel. This same condition applies to the energy radiated from *B* and reaching *A*. Thus the radiation in the plane of the elements,

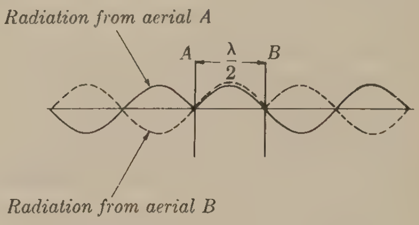


FIG.11-19 Radiation from two half-wave vertical elements spaced $\frac{\lambda}{2}$ apart.

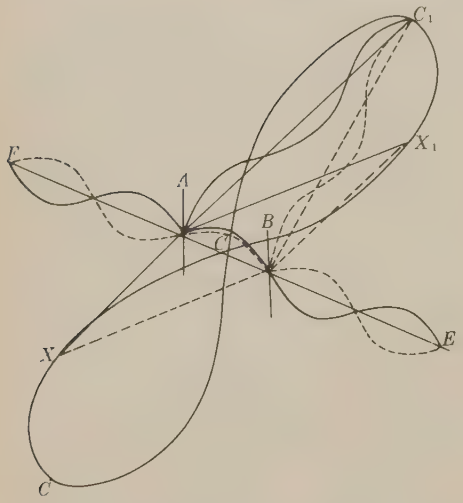


FIG.11-20 Radiation from two half-wave vertical elements spaced $\frac{\lambda}{2}$ apart.

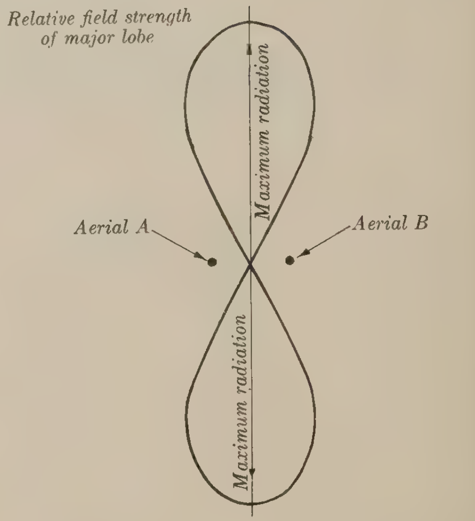


FIG.11-21 Horizontal directive pattern of broadside aerial.

that is the plane of the paper, is negligible. Consider the plane of the elements shifted as in Fig. 11-20. Along the directions ABE, BAF, as before, radiation is zero. In the direction CC_1 the distance from A to C is the same as from B to C and since the radiations start off in phase they reach point C in phase and are additive. Along any other direction XX_1 the distance from A to X differs from the distance from B to X. Thus an intermediate field strength is obtained at X and similarly at X_1. The directive pattern is as shown in the diagram with the maximum energy radiated in a direction broadside to the plane of the elements. If the elements A and B are looked down upon from a height and if the plane of the page represents the ground the horizontal directive pattern may be indicated as in Fig. 11-21. The surface pattern of the broadside aerial is shown in Fig. 11-22.

Broadside aerials are used with the elements erected vertically or horizontally. In the former case horizontal directivity is obtained while the vertical pattern is the same as for one element alone. In the latter case the pattern is sharpened in the vertical plane. The height required limits the number of elements which can be suspended horizontally so that more than two are not often used. The gain obtained from a broadside aerial is increased by increasing the number of elements. Fig. 11-23 shows the relative gain and directivity of broadside aerials with 2, 4 and 6 horizontal elements. As in the collinear aerial, radiation is bi-directional.

C. The Broadside Array

The *broadside array* consists of collinear elements stacked above and below

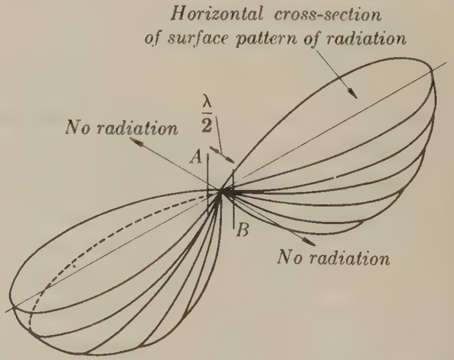


FIG.11-22 Surface pattern of broadside aerial.

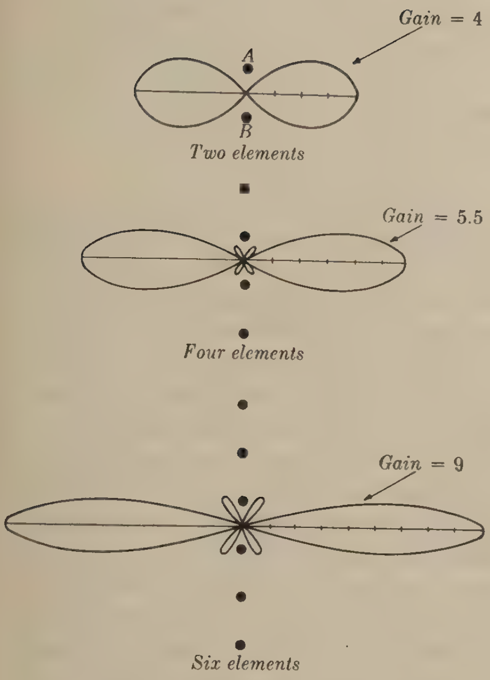


FIG. 11-23 Relative gain and directivity of broadside aerials with 2, 4 and 6 horizontal elements.

another set of similar elements as shown in Fig. 11-24. Thus two broadside aerials may be combined to form an array. The gain of broadside arrays depends upon the spacing, number and tuning of the elements. In practice high gains are readily obtainable. In these arrays radiation is concentrated into a narrow horizontal beam at very low angles above the horizon. The low-angle radiation is obtained as a result of the broadside aerial pattern.

The method of feeding the various elements of a broadside array may be understood by a consideration of Fig. 11-24.

The single pair of elements shown in Fig. 11-24 (a) are fed in phase since the feeder line makes the right end of each element positive at the same instant and by the same amount. This voltage condition also causes the currents in each element to be in phase and have the same magnitude.

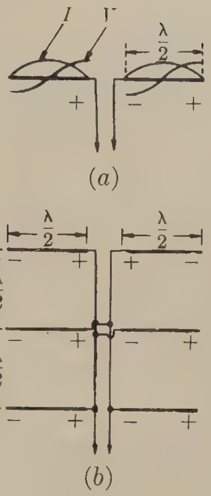


FIG. 11-24 Method of feeding elements of a broadside array.

The three-pair broadside array shown in Fig. 11-24 (b) is excited by connecting the pairs at intervals of $\frac{\lambda}{2}$ along the transmission line as indicated. At intervals of $\frac{\lambda}{2}$ along the line the voltage has opposite phases regardless of whether the line is resonant or non resonant. Further, the voltages are of the same magnitude at these intervals if the line is lossless and the load is concentrated at the end or at positions an integral number of half wavelengths from the end.

Fig. 11-25 shows a stacked dipole array consisting of 16 horizontal $\frac{\lambda}{2}$ elements, four erected in a line horizontally and four in parallel vertically. The method of feed is again accomplished by connecting at points which are at intervals of half wavelengths along the transmission line with reversed connections for each pair. This array radiates a narrow beam in both directions broadside to the plane of the array elements. If a still narrower beam is required addi-

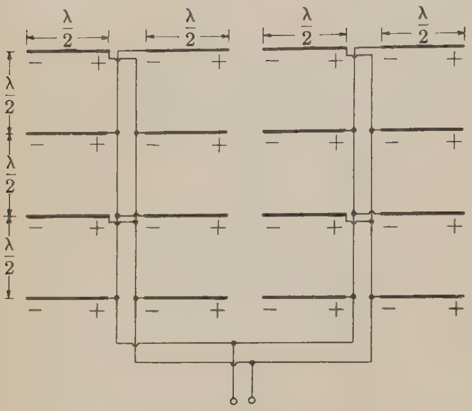


FIG. 11-25 Stacked dipole array consisting of 16 horizontal $\frac{\lambda}{2}$ elements, four in line horizontally and four in parallel vertically.

tional elements may be added both vertically and horizontally. The pattern may be made narrower in the vertical plane by stacking more elements one above the other; it may be made narrower in the horizontal plane by increasing the number of collinear elements.

From the above methods of feeding the elements of a broadside array it can be seen that the part of the transmission line within the array becomes a resonant line and there are standing waves on it. Thus this part of the line radiates energy but because of the close spacing and opposite phase of current in each line the fields cancel and the effect on the array pattern is negligible.

The practical method of reversing the element connections is to twist the line and so avoid any lack of symmetry which otherwise may be introduced.

D. The End-Fire Aerial

The simple *end-fire aerial* consists of two dipoles mounted parallel to each other, one quarter wavelength apart and driven so that their currents are 90° out of phase. The principle of the end-fire aerial may be understood by consideration of Fig. 11-26. Two half-wave elements are erected vertically in free space one quarter wavelength apart. Since the currents in A and B are equal and 90° out of phase the fields radiated from the elements at any instant are the same in amplitude and are 90° out of phase. By the time the radiated energy reaches B the energy leaving B towards C has the same phase and so they are additive. Along any other direction OX the distance from A to X in the horizontal plane differs from the distance from B to X. Thus an intermediate field strength is obtained at X. However, by the time the radiated energy leaving element B reaches A the energy leaving A is 180° out of phase causing complete cancellation in the direction OC_1. The directive pattern in the horizontal plane is shown in Fig. 11-26 and Fig. 11-27

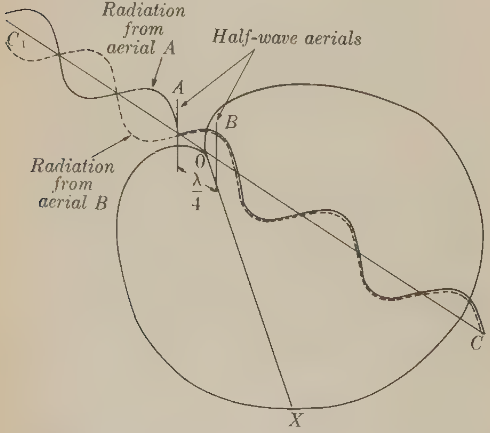


FIG. 11-26 Principle of end-fire aerial.

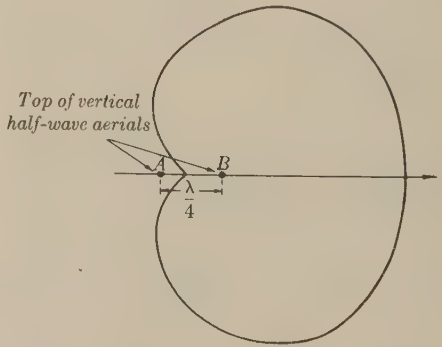


FIG. 11-27 Horizontal directive pattern of the end-fire aerial.

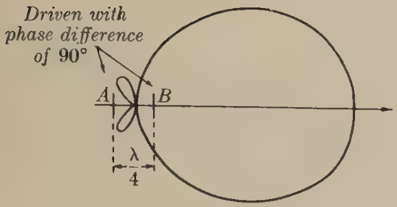


FIG. 11-28 Vertical directive pattern of the end-fire aerial.

with the maximum energy radiated in a direction along the plane of the elements towards the element *B*. The other element *A* is therefore sometimes referred to as a *driven reflector*.

In the vertical plane containing the elements there is a similar directional effect but it is modified by the fact that radiation from the ends of the elements is zero. Thus, in the vertical plane the beam is narrower and the directive pattern is shown in Fig. 11-28. The complete surface pattern is shown in Fig. 11-29.

The gain depends on the number of elements, spacing and phasing. The end-fire aerial gives both horizontal and vertical directivity regardless of whether it is mounted horizontally or vertically. If the elements are mounted horizontally a broad vertical pattern together with a sharper horizontal pattern results.

The plane of polarization of any directional system is the same as the plane of polarization of one of the elements. End-fire aerials provide the greatest gain and directivity for a given space. However, since their radiation resistance is low they are difficult to feed.

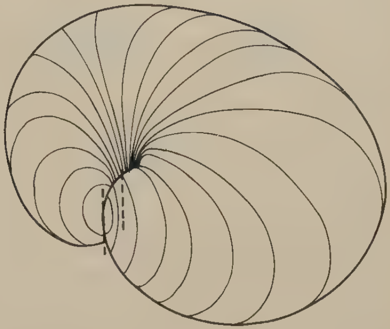


FIG. 11-29 Surface pattern of the end-fire aerial.

11.4 Parasitic arrays. If a wire or rod one half wavelength long is erected parallel to and near a driven dipole, it has a current induced in it and re-radiation from it results. The re-radiated energy combines with that directly radiated by the driven element and modifies the directional pattern of the dipole. The modifications in the pattern depend upon the relative positions of the two rods, the magnitudes of the currents in them and the phases of these currents. The free wire or rod is a parasitic element and is said to be *parasitically excited*. It is characteristic of parasitic-element systems to provide maximum radiation in one direction only as compared with the many directive systems which are bi-directional. This is a useful feature, especially in rotatable systems, because it helps to reduce interference due to reception from directions other than that along which communication is being carried on and also reduces the interference which the transmitter might otherwise cause in those same directions. The parasitic element, which is always erected parallel to the driven element, may be either,

(i) a *reflector* if it is erected on the side opposite to the major lobe,
(ii) a *director* if it is erected on the side containing the major lobe.

Figs. 11-30 and 11-31 indicate the relative positions of a director and a reflector respectively while Fig. 11-32 illustrates an 11-element aerial system using three reflectors and seven directors. This aerial system has reflectors above, below and behind the driven element and seven directors in the line of the major lobe.

A. The Reflector

Consider a parasitic element *B* one half wavelength long (Fig. 11-33) placed

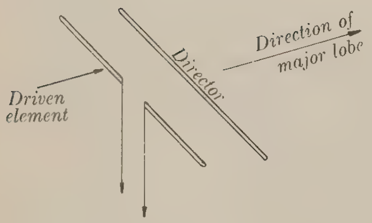


FIG.11-30 Dipole with director element.

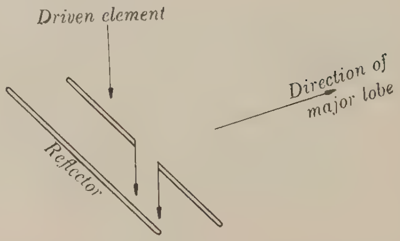


FIG.11-31 Dipole with reflector element.

parallel to and one quarter wavelength from a driven vertical element A. Since the parasitic is $\frac{\lambda}{4}$ from the driven element the field at the parasitic is 90° behind the field radiated at the aerial. A current is induced in the parasitic and this current, obeying Lenz's Law, tends to oppose the effect producing it. This means a phase shift of 180° is introduced and thus the energy re-radiated from the parasitic is 90° ahead of that radiated from the driven element. By the time the re-radiated energy reaches A it is exactly in phase with the energy radiated from A towards C and so the fields are additive. Along any other line OX in the horizontal plane the distance from A to X differs from the distance from B to X and an intermediate value of field strength is obtained at X. However, by the time the radiated energy travelling out from A towards C_1 reaches B it is exactly in anti-phase with the re-radiated energy from B towards C_1 causing cancellation of the fields in the direction OC_1. The directive pattern in the horizontal plane is similar to the pattern obtained from an end-fire array. Further, as for the end-fire array, there is a similar directional effect in the vertical plane containing the aerials but somewhat narrower since there is zero radiation from the ends of the aerials.

In practice it has been found that it is not necessary to have the reflector exactly one quarter wavelength behind the driven element. If the parasitic element is lengthened it becomes inductive. This produces a lag in the induced current and hence a phase lag in the re-radiated field which may be compensated for by moving

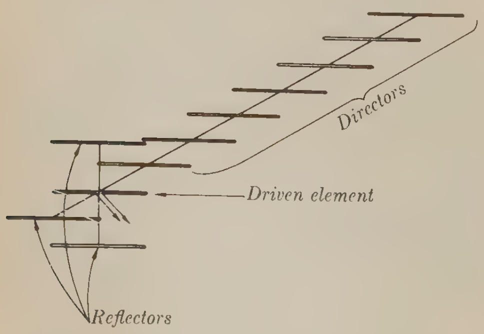


FIG.11-32 11-element aerial system using 3 reflectors and 7 directors.

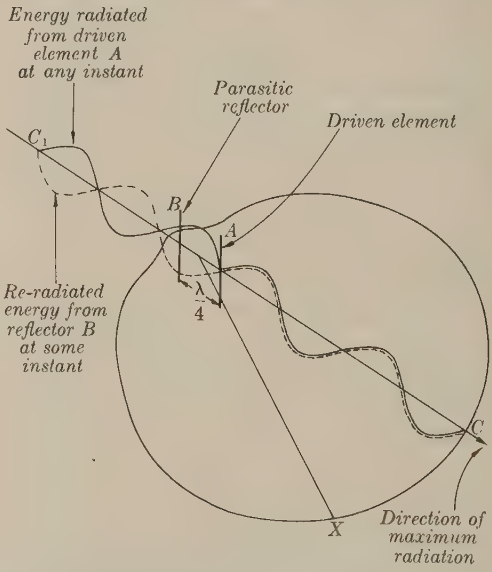

Energy radiated from driven element A at any instant
Parasitic reflector
Driven element
C1
B
A
λ/4
Re-radiated energy from reflector B at some instant
C
X
Direction of maximum radiation

FIG.11-33 Function of reflector.

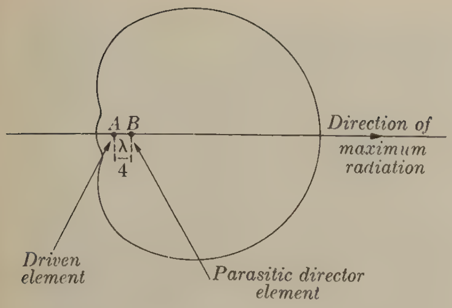


FIG. 11-34 Function of director.

the parasitic element closer to the driven element. Optimum results are obtained from a reflector slightly longer than one half wavelength located at a distance of 0.15 of a wavelength behind the driven element.

B. The Director

Consider a parasitic element *B* slightly less than one half wavelength long (Fig. 11-34) erected parallel to and one quarter wavelength from a driven vertical element *A*. Since the length of the aerial is now slightly less than one half wavelength it is capacitive. The change from inductive to capacitive reactance introduces an additional phase shift. The fields which cancelled in the case of the reflector now become additive and those which reinforced each other now cancel each other. Thus the maximum energy is now radiated in the direction of the parasitic element. The parasitic element has become a director and the horizontal directive pattern is shown in Fig. 11-34. Optimum results are obtained from a director which is slightly shorter than one half wavelength located at a distance of 0.1λ from the driven element.

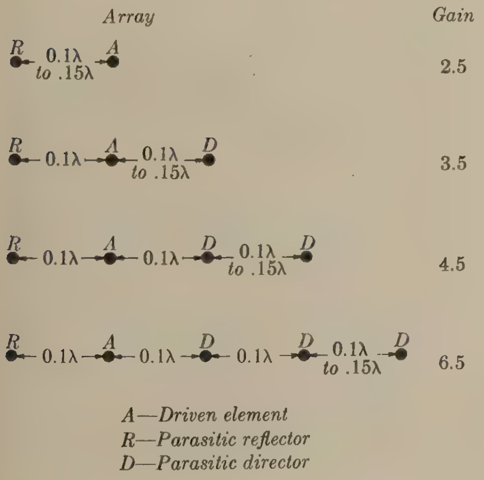


FIG. 11-35 Theoretical gain of aerial systems using directors and reflectors.

C. Multi-Element Parasitic Arrays

1. THE YAGI ARRAY.

Several parasitic elements may be used in conjunction with a driven element to further increase the directivity and gain. Combinations of driven element, reflector and director, together with the approximate theoretical gain are shown in Fig. 11-35. The directive pattern shown in Fig. 11-36 which is produced by such arrays demonstrates how the parasitic elements concentrate the radiation into a narrow uni-directional beam.

One common type of Yagi array is made up of ten directors, three reflectors and a driven folded dipole as shown in Fig. 11-37 (a) and (b). The reflectors above and below the driven element serve to cut off radiation in these directions and so reduce the width of the beam in the vertical plane. Fig. 11-38 (a) and (b) represent the vertical and horizontal directive patterns when the elements of the array are mounted horizontally. Small lobes exist to the sides and rear of the major lobe.

A smaller Yagi aerial system consisting of a folded dipole, three directors,

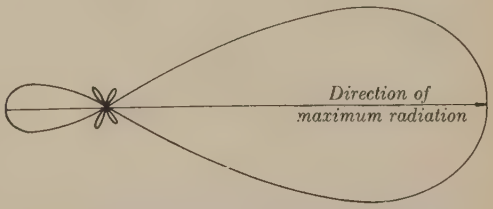
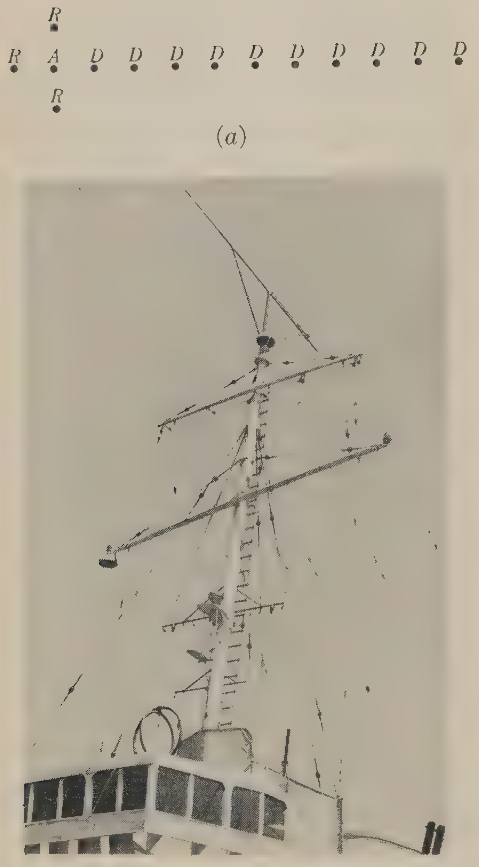


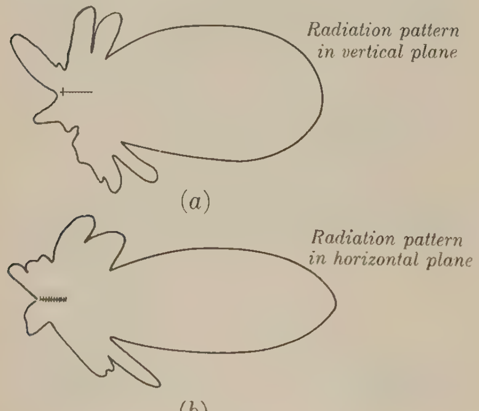
FIG. 11-36 Directive pattern of Yagi array.



(a)

(b)

Fig.11-37 Yagi array of 10 directors and 3 reflectors.



(a)

(b)

Fig.11-38 Vertical and horizontal directive pattern of a horizontal Yagi of 10 directors and 3 reflectors.

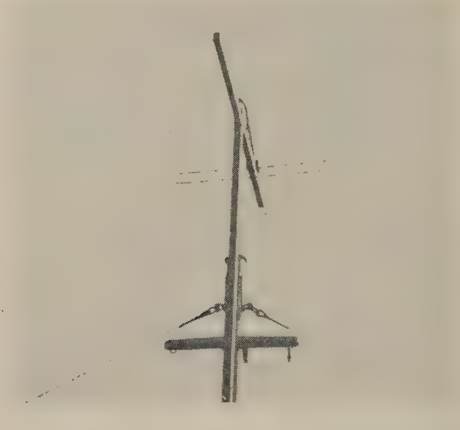


Fig.11-39 Yagi array of 3 directors and 1 reflector.

and 1 reflector is shown in Fig. 11-39.

A larger Yagi array often referred to as the *fishbone* aerial is shown in Fig. 11-40 and consists of six Yagi units. In this case the reflectors above, below and behind the driven element of the regular Yagi are replaced by a solid reflecting surface. Directivity in the plane containing the element supports is obtained because of the length of the radiating source as in a broadside array. Directivity in a plane perpendicular to the plane containing the element supports is obtained as a result of the directors and solid reflecting surface.

2. The Lazy-H Array.

The lazy-*H* array is a combination of driven and parasitic elements. It consists of four centre-fed dipoles grouped in the form of an *H* as shown in Fig. 11-41. Each of the driven dipoles has a parasitic element mounted behind it, properly spaced and cut to act as a reflector. This array combines the directional effects of the collinear array, broadside array and parasitic reflector array. In practice the dipoles are mounted horizontally and the directivity in the horizontal plane is the same as the directivity of a collinear array with reflectors as shown in Fig. 11-42. The

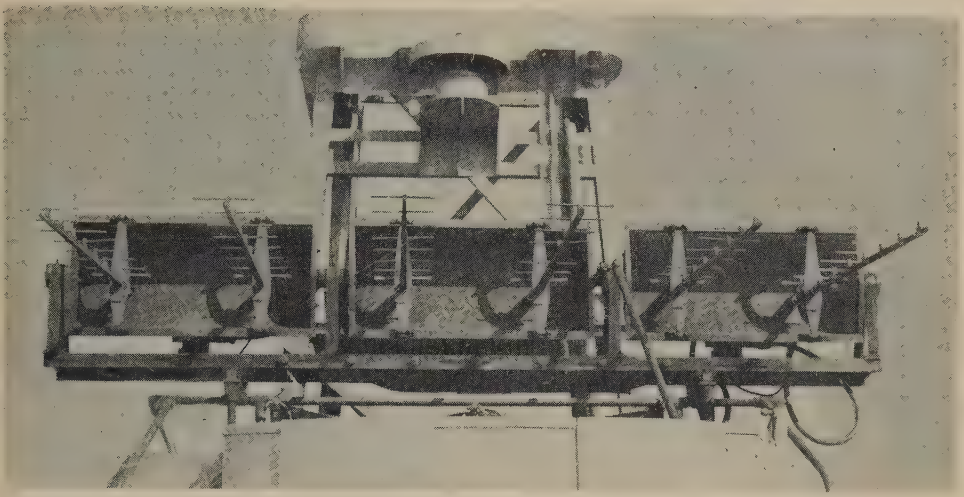
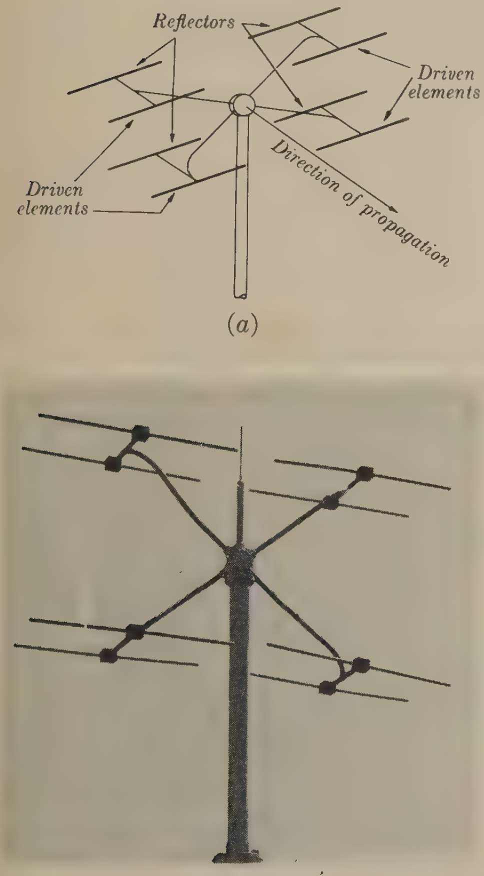


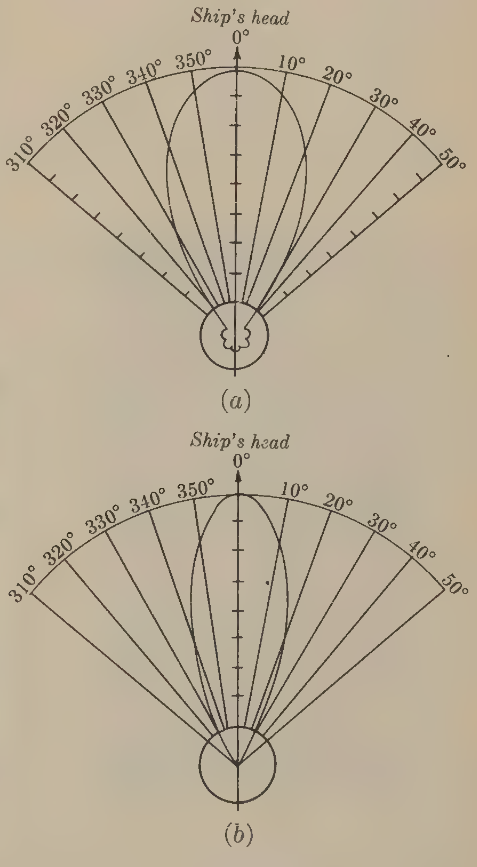
FIG.11-40 Array of 6 Yagi units sometimes referred to as the fishbone aerial.



(a)

(b)

FIG.11-41 The Lazy-*H* array.



(a)

(b)

FIG.11-42 Horizontal radiation pattern of the Lazy-*H* array.

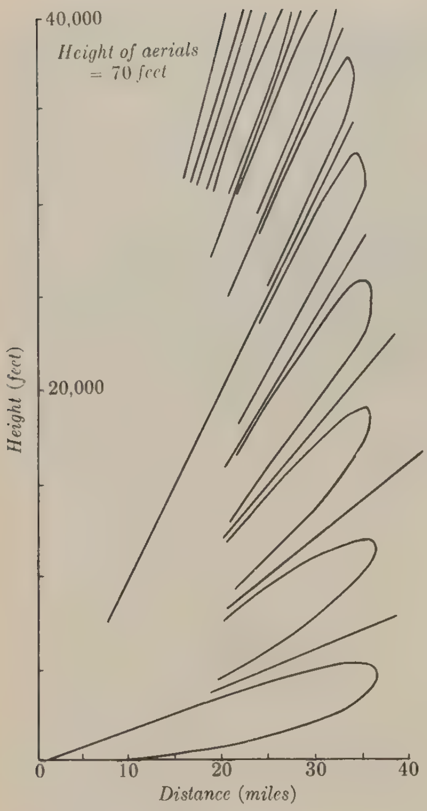


FIG.11-43 Vertical radiation pattern of Lazy-*H* array.

directivity in the vertical plane is that provided by a broadside array (Fig. 11-43). The side lobes for the lazy-*H* array are smaller than those of the Yagi array.

3. THE BROADSIDE CURTAIN OR BILL-BOARD ARRAY.

The broadside array which provides bi-directional radiation is of no use in radar applications. One of the main lobes is eliminated by either of two methods:

(i) by means of a second similar broadside array properly phased and spaced with respect to the first
(ii) by means of a reflecting screen.

If the first method is used the best arrangement results when the second array is placed $\frac{\lambda}{4}$ behind the first. The

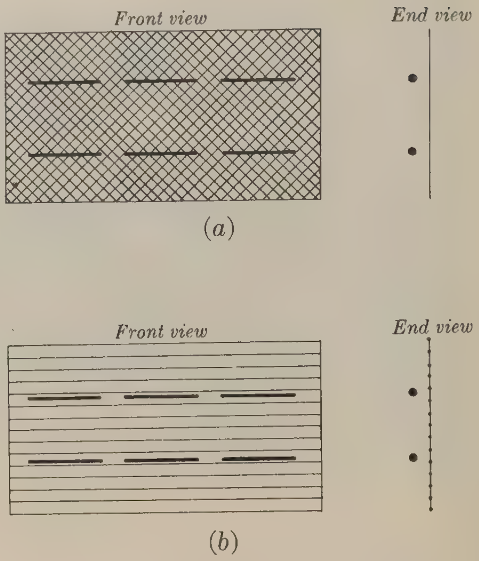


FIG.11-44 The broadside curtain or billboard array.

second array is excited with currents which are equal in magnitude but which lead the main array currents by 90°. In practice, the added broadside array usually consists of parasitic elements. These elements are not placed exactly $\frac{\lambda}{4}$ behind the driven elements as is the case for parasitic reflectors.

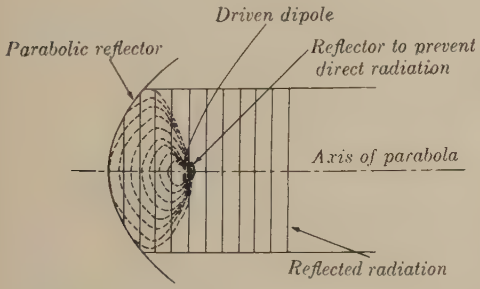
The second method of producing uni-directional radiation from a broadside array is to place a conducting screen or curtain approximately $\frac{\lambda}{4}$ behind the array. This conducting curtain (Fig. 11-44) consists of a screen or grid structure to reduce weight and wind resistance and eliminate any visibility difficulties. The screen of Fig. 11-44 (a) permits the array elements to be placed either vertically or horizontally while the screen shown in Fig. 11-44 (b) imposes the conditions that the array elements be parallel to the screen wires. The presence of a screen eliminates back radiation by reflection and at the same time it sharpens the forward pattern.

11.5 Parabolic reflectors. The fundamental principles which apply to aerials at the lower frequencies also apply at the ultra-high frequencies. Microwave directive aerials may be built into a very small space. At the microwavelengths parabolic and horn type reflectors, which are too large and bulky at the lower frequencies, find wide application.

At the very high frequencies electromagnetic waves and light waves become identical in character. Thus the reflector property of the parabola may be utilized at the microwavelengths. If a dipole is placed at the focus of a parabola (Fig. 11-45) there is current induced in the copper parabola which re-radiates a new field concentrated into a beam in a direction parallel to the axis of the parabola. Direct radiation from the driven dipole is usually prevented by means of a flat or hemispherical reflector so that interference effects between the direct radiation and the reflected radiation are minimized.

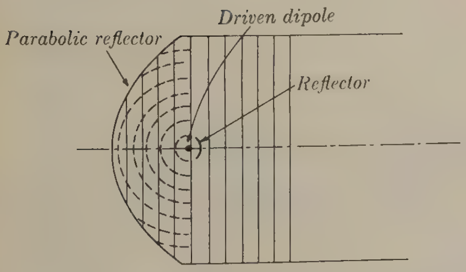
One type of practical parabolic reflector or mirror, shown in Fig. 11-46, consists of a parabolic spheroid which resembles a saucer. It is sometimes referred to as a *parabolic dish.* Very narrow beams are possible with this type of reflector. Typical directional patterns obtained from a parabolic dish reflector are shown in Fig. 11-47 (a) and (b).

Fig. 11-48 shows a parabolic reflector which is made up of rods. This serves to reduce wind resistance which is appre-



Horizontal section

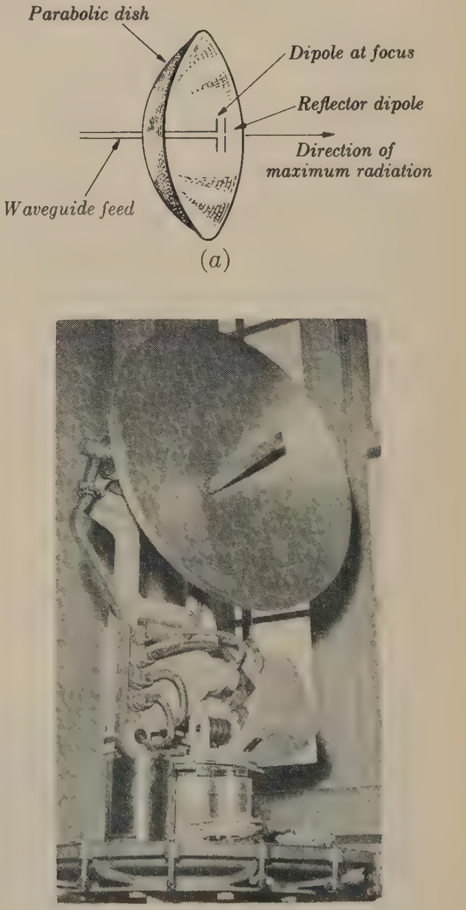
(*a*)



Vertical section

(*b*)

Fig. 11-45 Operation of parabolic reflector.



(*b*)

Fig. 11-46 3 cm. dipole with parabolic reflector.

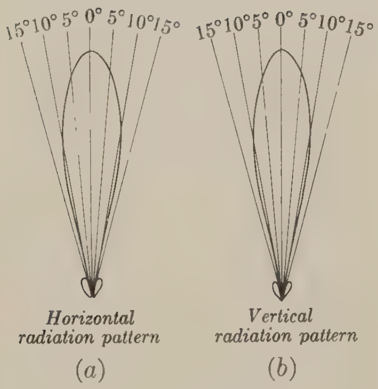


Fig. 11-47 Directional patterns obtained for a parabolic disk reflector.

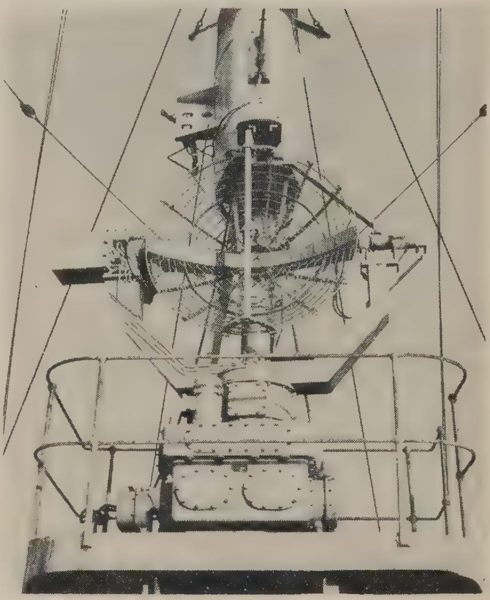


Fig. 11-48 Parabolic basket.

ciable in a parabolic reflector of this physical size. This reflector is often referred to as a *parabolic basket.*

Another type of parabolic reflector is the *parabolic slice* shown in Figs. 11-49, 11-50, 11-51 and 11-52. Figs. 11-49 and 11-50 show a parabolic slice with closed top and bottom while Figs. 11-51 and 11-52 shows one with an open top and bottom. In each case the radiating element is located at the focus on the axis of the parabola. Slices are usually designed so that the focus lies well within the mouths so that almost all the radiated energy is intercepted and reflected by the parabolic surfaces.

11.6 Feeds for parabolic aerials. The feed system for a parabolic aerial may employ coaxial lines or waveguides and may be classified as *rear* or *front* feed.

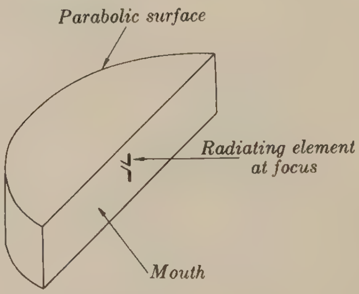


Fig. 11-49 Parabolic slice with closed top and bottom.

Fig. 11-50 Parabolic slice with closed top and bottom.

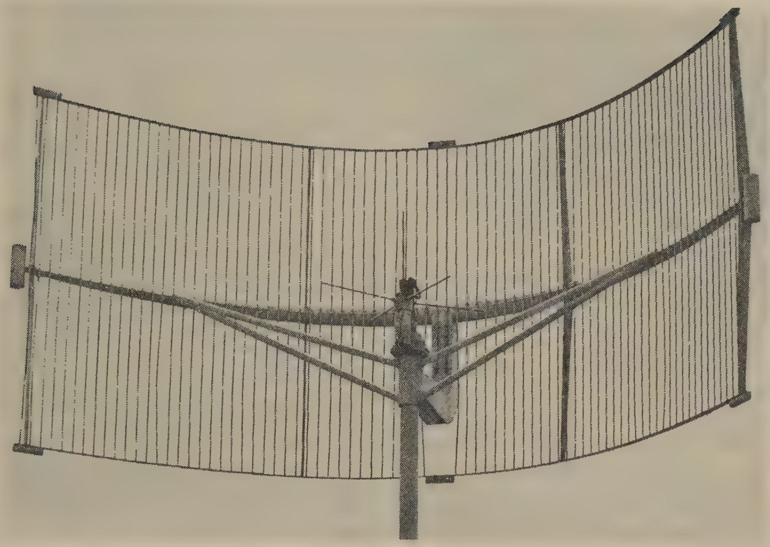


FIG.11-51 Parabolic slice with open top and bottom.

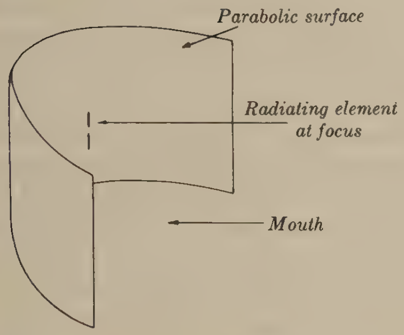


FIG.11-52 Parabolic slice with open top and bottom.

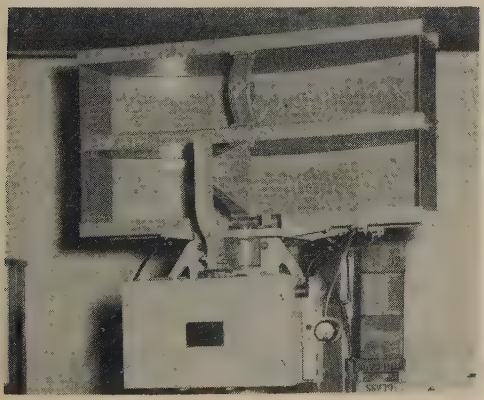


FIG.11-53 Front waveguide feed of a parabolic slice.

A. Front Waveguide Feeds

A front waveguide feed of a parabolic slice is shown in Fig. 11-53. In front of, and slightly below the transmitting aerial mirror is the magnetron oscillator box. Projecting upwards out of the oscillator box is the lower part of the waveguide feed, which is in line with but separated from the upper part of the waveguide feed which is fixed to the parabolic slice. The gap between the two parts fulfills two purposes.

(i) It isolates the rigid upper part from the shock-absorber-mounted lower part (oscillator box).

(ii) It allows the magnetron anode, connected electrically to the lower part of the waveguide, to be insulated from earth, so that a current transformer can be inserted between the anode and earth in order to observe the anode current.

The energy radiated along the waveguide is projected from the flared end of the guide onto the parabolic slice from which surface it is reflected. This aerial system is used in radar surface-search

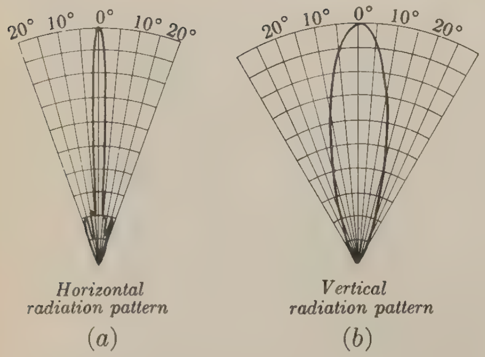


Fig. 11-54 Radiation pattern of aerial system of Fig. 11-53.

systems which require a highly directive beam in the horizontal plane and less directivity in the vertical plane because of the rolling and pitching of the ship. The radiation pattern is shown in Fig. 11-54. The flared section of the waveguide, sometimes called a *horn*, is oriented to produce vertical polarization. The flare increases the directivity and also the gain by reducing the amount of primary radiation that escapes reflection. The length of the horn and the flare angle determines the directivity of the radiation which energizes the reflector and also affects the terminating impedance for the waveguide. Flaring the end of a waveguide results in a decrease in guide wavelength so that the wavelength at the mouth of the horn can be made nearly the same as the free-space wavelength. At the same time, flaring the end of the waveguide also results in a change in the wave impedance at the mouth of the horn to a value which approaches the wave impedance in free space. Thus such a horn can be used to radiate into free space as the proper termination of the guide for maximum transfer of energy is realized. Also when used to feed a parabolic reflector satisfactory terminations may be obtained by adjustment of the horn dimensions and the spacing between the horn and reflector.

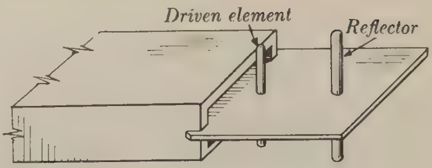


Fig. 11-55 Rear waveguide feed.

The parabolic slice of Fig. 11-50 is also fed by means of a flared waveguide.

B. Rear Waveguide Feed

The 3 cm. aerial system shown in Fig. 11-46 employs *rear waveguide feed.* An expanded diagram of the feed is shown in Fig. 11-55. The dipole radiating element is so placed that nearly all the energy from the waveguide is absorbed and re-radiated from the aerial. This is partly accomplished by means of the second dipole reflector placed one quarter of a wavelength from the driven dipole.

A rear waveguide feed which finds frequent application is the *Cutler* feed shown in Fig. 11-56. The waveguide is terminated in a rectangular box which may be considered as two small waveguides formed by dividing the main waveguide feed into two parts and bending each part back through 180°. The open slots are one half wavelength apart and radiate the energy toward the parabolic reflector. Since the slots are one half wavelength

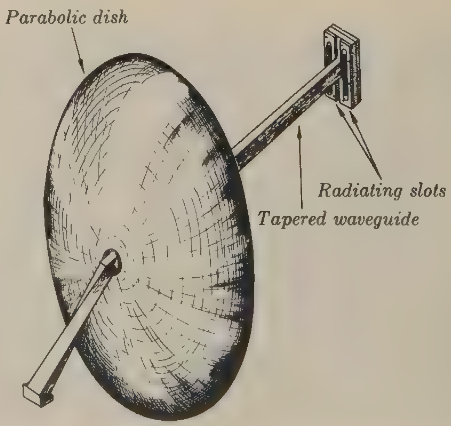


Fig. 11-56 Cutler rear waveguide feed.

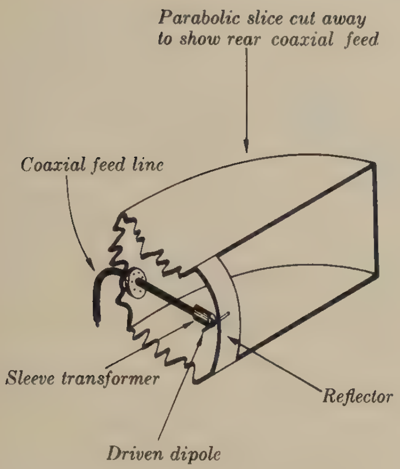


FIG.11-57 Coaxial feed to 10 cm. radar aerial.

apart the shorter dimension of the waveguide must be very small at the terminal point where the waveguide is divided and bent back. To provide for this the main wave guide is tapered and mismatch, which would otherwise be introduced by a sudden change in the shorter dimension of the waveguide is avoided. Tapering the guide in the shorter dimension does not affect the wave impedance of the guide (Sec. 10.24).

C. Coaxial Feeds

Fig. 11-57 shows a coaxial feed for a receiving aerial system of 10 cm. radar equipment. The receiving mirror receives energy reflected from the target and concentrates it at the dipole placed at the focus of the mirror. The dipole is set in a polythene cap and each arm of the dipole is 2.25 cm. which is less than one-quarter of 10 cm. since the wavelength in polythene is less than in air. The overall dipole length is 4.9 cm. A small parabolic slice, fixed in front of the dipole, reflects back onto the dipole any energy from the mirror which has passed the dipole, thus increasing the efficiency of the system and preventing direct re-radiation toward the target. The dipole is connected via a sleeve transformer to the unbalanced coaxial line (Fig. 11-58). The concentric line is taken through the back of the mirror and so to the receiver.

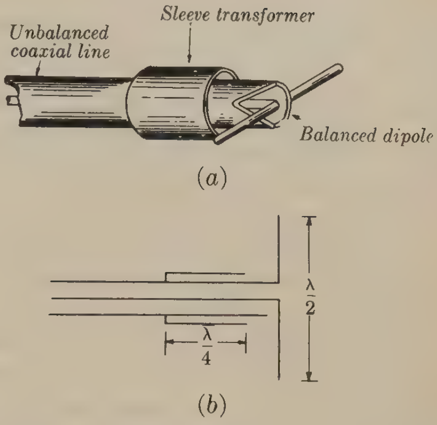


FIG.11-58 Dipole connected via sleeve transformer to an unbalanced coaxial line.

Expanded diagrams of other coaxial feeds used with 10 cm. radar equipment are shown in Fig. 11-59. In these dia-

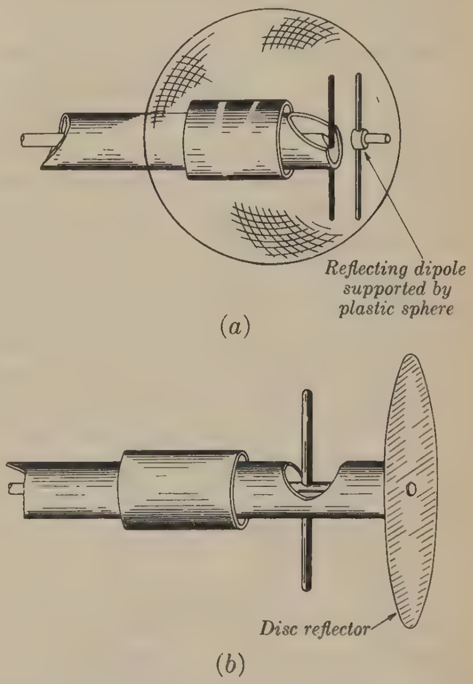


FIG.11-59 Typical coaxial feeds used in 10 cm. radar equipments.

grams only the dipole sections of the aerial systems are shown with parabolic mirrors omitted. Fig. 24-59 (a) shows a dipole connected to a $\frac{\lambda}{4}$ sleeve transformer and a dipole reflector. Fig. 24-59 (b) employs a phase-disc reflector.

11.7 Horn radiators. *Horn radiators* are of sufficiently small physical size to become practical at the microwavelengths. These radiators do not involve resonant elements and consequently they are usable over a wide frequency band. The operation of the horn radiator is similar to the horns of acoustic systems. The acoustic horn usually has a throat with dimensions smaller than the sound wavelengths for which it is used, while the throat of the electromagnetic horn has dimensions comparable to the wavelengths used. Horn radiators are usually fed by means of waveguides. The shape of the horn along with the dimensions of its mouth determines the field pattern. In general, the larger the opening of the horn the more directive is the resulting field pattern. An aperture of approximately five wavelengths produces a radiated major lobe of approximately 30°.

The important feature of a horn radiator is that a horn radiates a directive beam similar to that of an array, with reflectors, which covers an area approximately equal to the area of the mouth of the horn. The beams produced by horn radiators are shown in Fig. 11-60 together with the corresponding curtain arrays which produce the same pattern. Thus it is evident that the use of a flared waveguide or horn as a radiator simplifies the task of tuning and matching feeders in an elaborate curtain array.

When the end area of a horn radiator is reduced side lobes are produced. The directive pattern is a function of the shape and area of the open end of the horn.

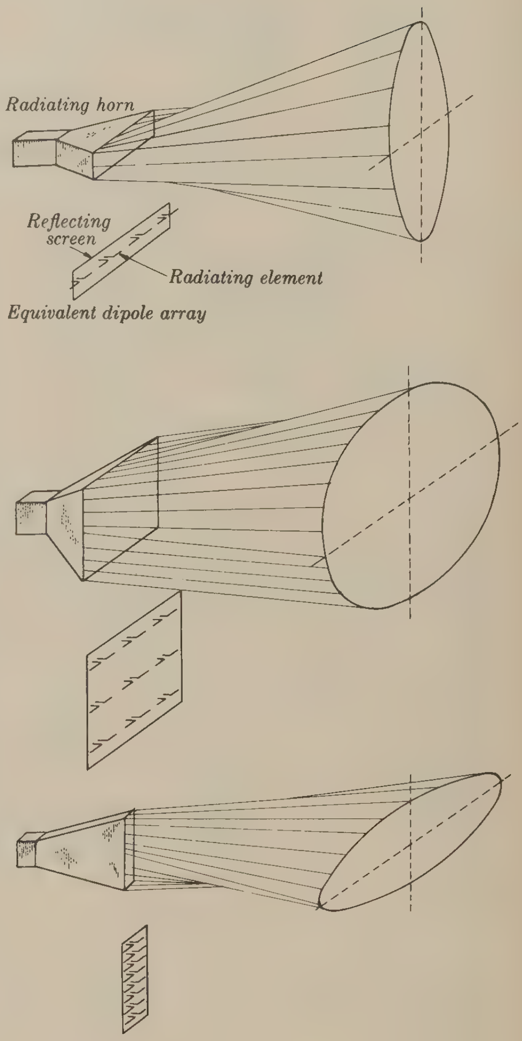


Fig. 11-60 Radiation from horn radiators.

11.8 Corner reflectors. The *corner reflector* consists of two flat conducting sheets that meet at an angle to form a corner (Fig. 11-61). This reflector usually has a driven dipole mounted so as to bisect the angle at the corner. Maximum radiation is in the horizontal plane. The corner reflector provides greater gain than the parabolic reflector. Greatest gain is obtained when the distance *d* is a half wavelength and the angle between the sides is 45 degrees. The sheets are made the same height as the dipole and

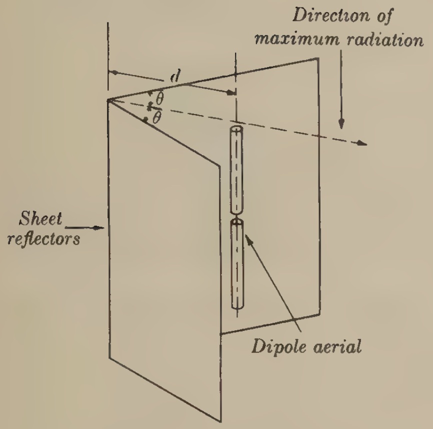


FIG.11-61 Corner reflector.

the width of a sheet approximately two wavelengths. The sheet width is not critical.

11.9 Ground effects. The directional performance of an aerial is modified considerably by the presence of the earth underneath it. All of the directional patterns thus far discussed have been for aerials in free space. At ordinary frequencies the distance between the aerial and earth plays an important part. The earth acts like a huge reflector for those radiated waves lower than the horizontal. The angle of reflection is the same as the angle of incidence (Fig. 11-62). The reflected waves combine with the direct waves, or those radiated at angles above the horizontal in various ways depending upon the manner in which the aerial is oriented, its height, its length and the character of the ground.

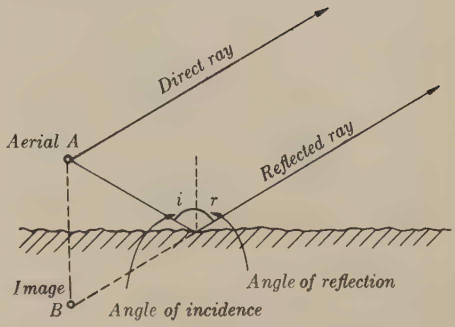


FIG.11-62 Ground effect.

At some vertical angles above the horizontal the direct and reflected waves are exactly in phase and the resultant field is equal to the sum of the two fields.

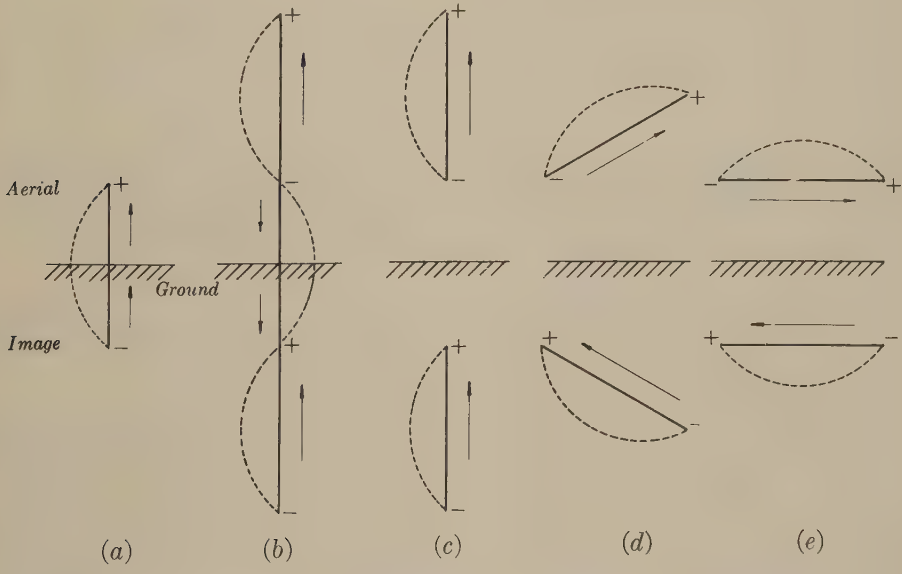
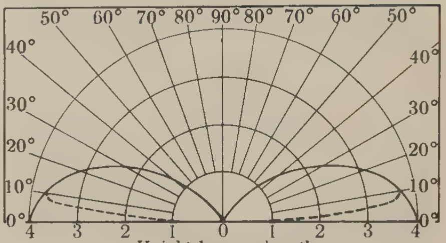


FIG.11-63 Image aerials indicating relative polarity to actual aerial.

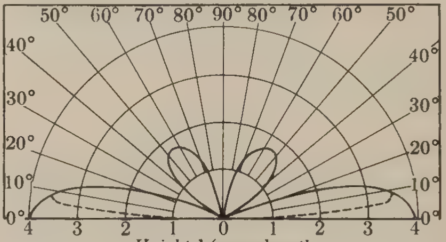
At other vertical angles the two waves are completely out of phase and the resultant field is the difference between the two. At still other angles the resultant field will have intermediate values. Thus the net effect of the ground is to increase the intensity of radiation at some vertical angles and decrease it at other angles.

It is often convenient to think of the reflected ray as appearing to come from an underground image of the real aerial (Fig. 11-62). The image is always as far below the ground as the real aerial is above it. Like an image in a mirror, the image aerial is in reverse as shown in Fig. 11-63. If the real aerial is horizontal, Fig. 11-63 (e), and it is instantaneously charged so that one end is positive and the other end negative as shown, then the image aerial which is also horizontal is oppositely poled. Similarly if the lower end of a half-wave vertical aerial, Fig. 11-63 (c), is negative at any instant, the end of the vertical image nearest the surface is positive. Thus the currents in the horizontal aerial and its image are 180° out of phase but the currents in the vertical aerial and its image are in phase. The effect of ground reflection, or the image aerial, is therefore different for horizontal and vertical dipoles.

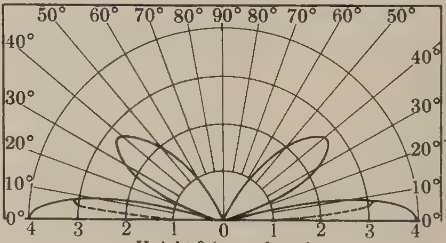
Thus the relative intensity of radiation at various vertical angles will depend upon the free-space radiation pattern of the aerial alone and the modifications of that pattern as a result of reflection of the waves from the ground. The vertical-plane radiation patterns for a vertical half-wave aerial above perfectly conducting ground are shown in Fig. 11-64. The height is that of the centre of the aerial. The lower end of the actual pattern is curved inward as shown by the dotted lines, due to attenuation of the low-angle radiation because of ground losses.



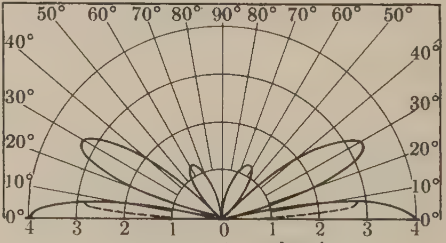
Height ¼ wavelength



Height ½ wavelength



Height ¾ wavelength

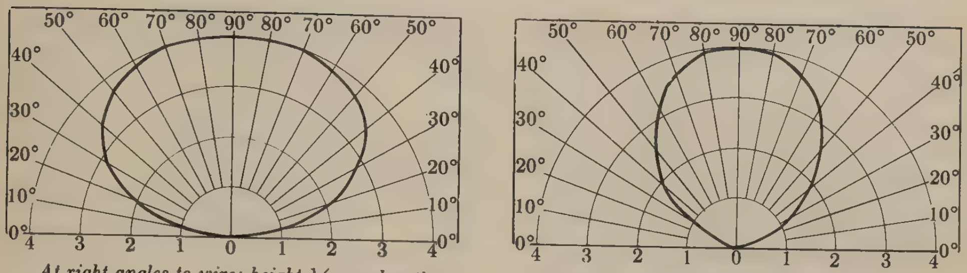


Height 1 wavelength

Fig. 11-64 Vertical-plane radiation patterns for a vertical half-wave aerial above perfectly conducting ground.

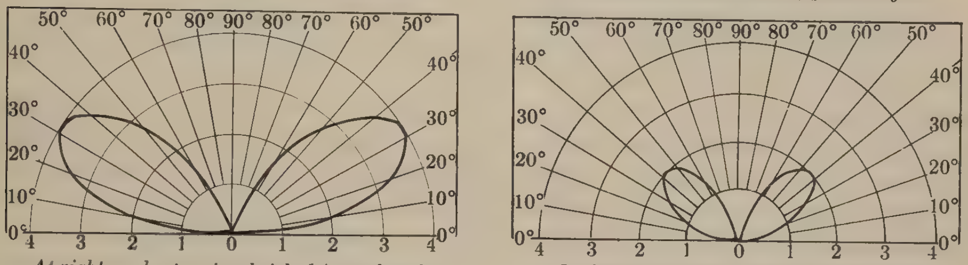
The vertical-plane radiation patterns for a horizontal half-wave aerial above perfectly conducting ground are shown in Fig. 11-65.

From Figs. 11-64 and 11-65 it may be seen that the additions of the direct and reflected fields in some directions and subtractions in other directions cause the



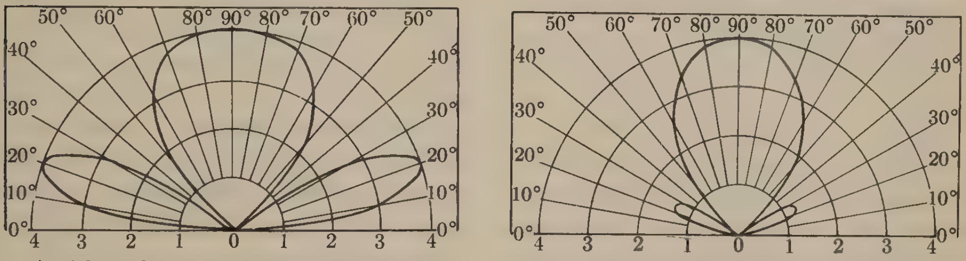
At right angles to wire; height ¼ wavelength

In direction of wire; height ¼ wavelength



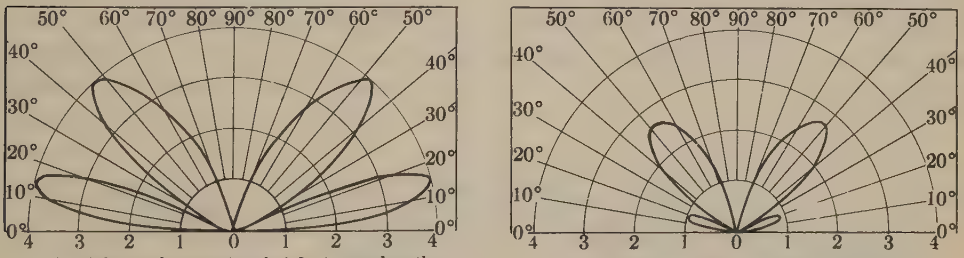
At right angles to wire; height ½ wavelength

In direction of wire; height ½ wavelength



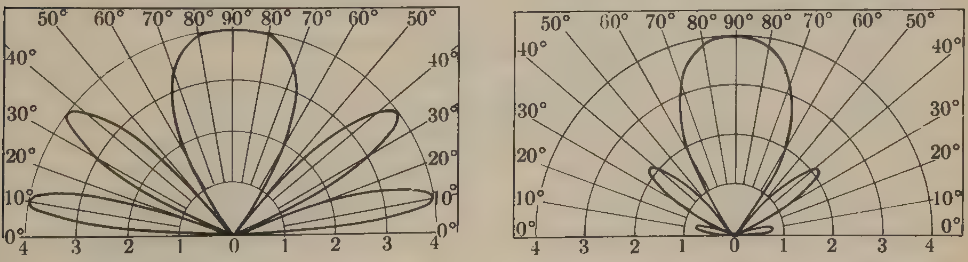
At right angles to wire; height ¾ wavelength

In direction of wire; height ¾ wavelength



At right angles to wire; height 1 wavelength

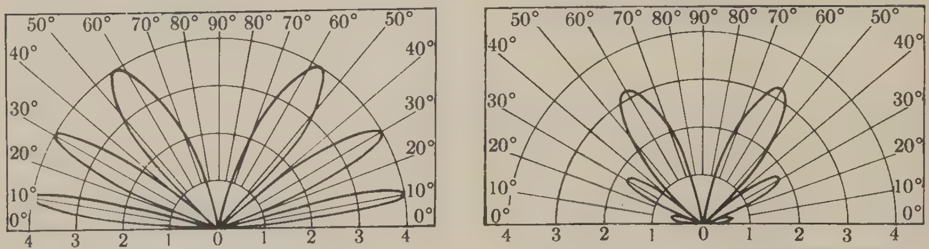
In direction of wire; height 1 wavelength



At right angles to wire; height 1¼ wavelength

In direction of wire; height 1¼ wavelength

FIG. 11-65 Vertical-plane radiation patterns for a horizontal half-wave aerial above perfectly conducting ground.



At right angles to wire; height 1½ wavelength

In direction of wire; height 1½ wavelength

Fig.11.65 Vertical-plane radiation patterns for a horizontal half-wave aerial above perfectly conducting ground.

aerial to have a non-uniform field pattern. Thus, where reinforcement occurs lobes are present and where cancellation occurs gaps appear. The main factors which determine the angle of elevation of the lobes and gaps in the vertical polar diagrams are the wavelengths and the height of the aerial above ground. If the height of the aerial is large compared to the wavelength the first lobe is at a very low angle.

The vertical free-space pattern of a Yagi array is shown in Fig. 11-38 (a). If this array is located near the ground this pattern is modified due to ground reflections to obtain a true picture of the directive field pattern. The modified field pattern of this directional array is shown in Fig. 11-66.

The effect of a perfectly conducting ground can be simulated, in the vicinity of an aerial, by means of a metal screen or mesh underneath the aerial. Such a screen often improves the performance of the aerial by reducing losses. The screen is most effective at high frequencies. It should extend at least one half wavelength in all directions from the aerial.

11.10 The quarter-wave aerial. A very important example of ground reflection occurs in the *quarter-wave* or *Marconi aerial.* This is a vertical quarter-wave aerial grounded at one end as shown in Fig. 11-67. This aerial acts like one half of a half-wave aerial and the ground acts as a mirror to supply the missing quarter-wave section. The current maximum is

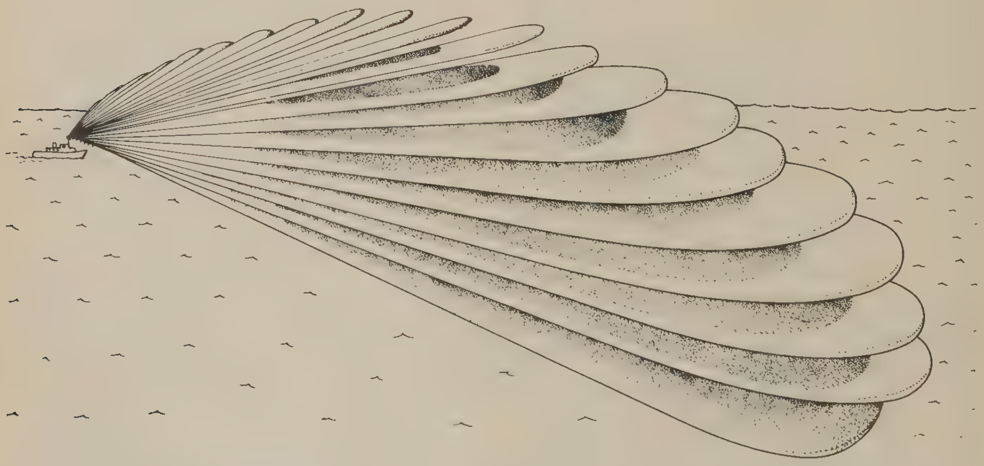


Fig.11-66 Directive field patterns of Yagi modified as a result of ground reflections.

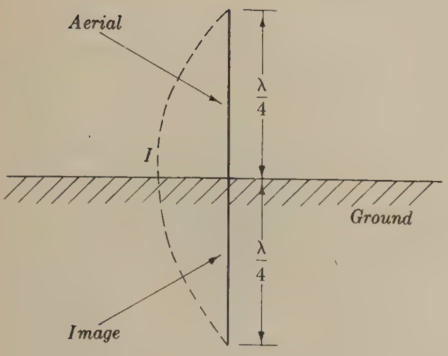


FIG. 11-67 Vertical quarter-wave (Marconi) aerial.

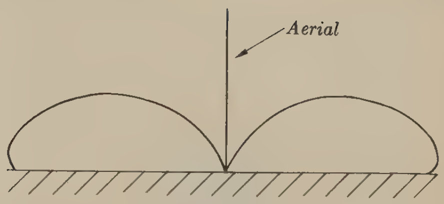


FIG. 11-68 Vertical directive pattern of vertical quarter-wave aerial.

at the base. The vertical directive pattern, Fig. 11-68, is thus the upper half of that for a dipole in free space.

11.11 The ground-plane aerial. One example of a *ground-plane aerial* is the *steering-wheel* or *life-saver* aerial, (Fig. 11-69). This aerial has a height of about 20 inches and diameter of 20 inches. The radiating conductor is a vertical quarter-wave rod which sends out vertically polarized waves. Below this radiator are a bakelite insulator and a horizontal wheel. The wheel portion of the aerial furnishes, for a limited distance, a ground plane so that the behaviour of the aerial is a good deal like that of a vertical quarter-wave aerial above a real ground. The aerial provides low-angle radiation acting to concentrate the energy toward the horizon like a simple vertical dipole. In effect the operation is similar to that of a vertical dipole and the only physical difference is that the lower part of the dipole is replaced by the horizontally mounted ring. In the horizontal plane the aerial has a circular radiation pattern so that the radiating efficiency in the various directions of azimuth is uniform unless it is disturbed by the presence of nearby metal structures. Good performance requires the aerial to be located as high and as much in the clear as practicable.

A second aerial employing a vertical radiator and simulated ground plane is shown in Fig. 11-70.

The undesirable effect of ground resistance which absorbs energy from the charging currents is minimized by the array of conductors at the base of the aerial. These conductors, sometimes called a *counterpoise*, provide a low-impedance path for the charging currents. The counterpoise rods are made approximately as long as the aerial is high.

11.12 Long single-wire aerials.

A. Resonant Long Single-Wire Aerials

As shown in section 11.1 the maximum radiation from a dipole is broadside to it. If the wire is now made a wavelength or more in length the radiation tends to concentrate more and more off the ends. As the number of half wavelengths is increased the radiation becomes more

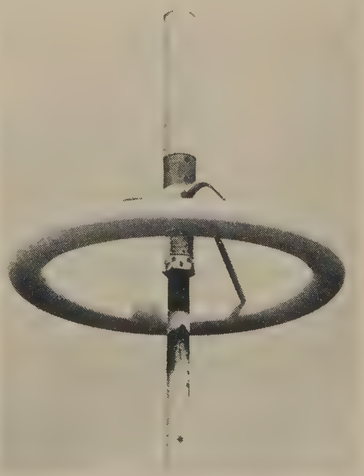


FIG. 11-69 Ground-plane (steering-wheel) aerial.

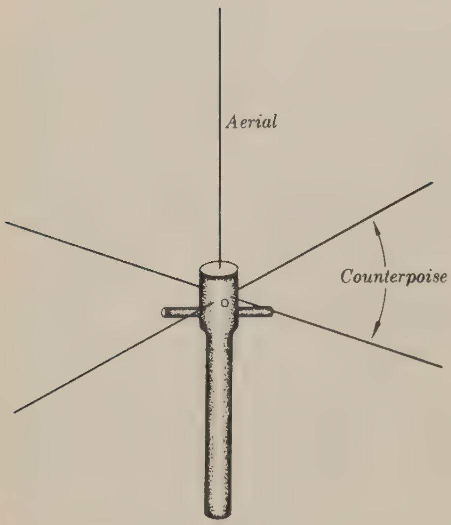


Fig.11-70 Vertical radiator with simulated ground plane.

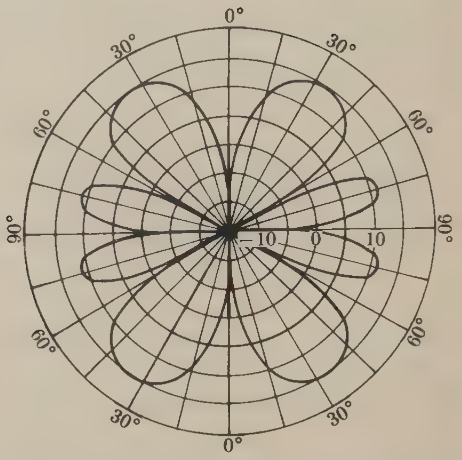


Fig.11-71 Horizontal directive pattern of a horizontal two-wavelength aerial in free space.

directive. Fig. 11-71 is the horizontal directive pattern of a horizontal two-wavelength aerial in free space. It should be noted that the major lobes occur at angles of about 36°. The long-wire aerial radiates more power in its most favourable direction than does a half-wavelength aerial in its most favourable direction. The power gain of course is obtained at the expense of radiation in other directions. Thus by properly locating a long-wire aerial it is possible to send out a stronger signal in a desired direction than is possible with a half-wave aerial. Practical directive patterns of horizontal long-wire aerials which give the relative field strengths for various vertical angles of radiation are shown in Figs. 11-72, 11-73, 11-74, 11-75.

The problems of feeding a long-wire aerial are similar to those when feeding

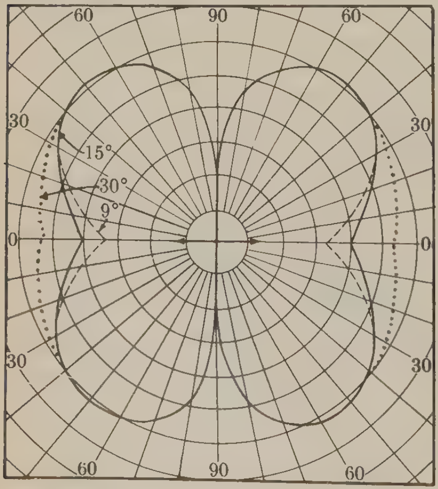


Fig.11-72 Horizontal directive pattern for 1-wavelength aerial at vertical angles of 9°, 15° and 30°.

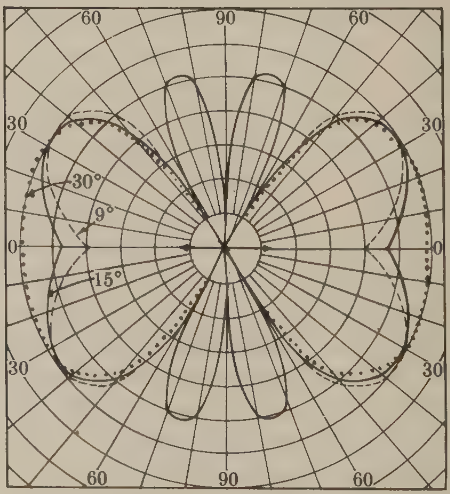


Fig.11-73 Horizontal directive pattern for two-wavelength aerial at vertical angles of 9°, 15° and 30°.

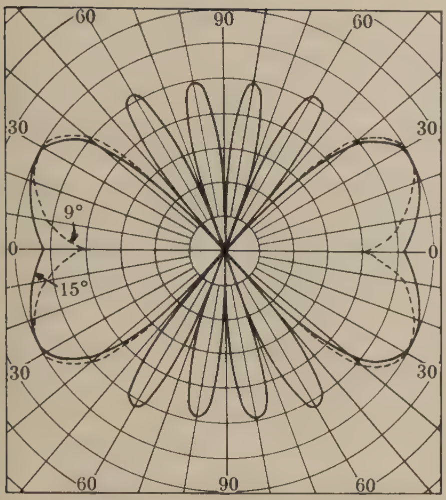


FIG. 11-74 Horizontal directive pattern for three-wavelength aerial at vertical angles of 9° and 15°.

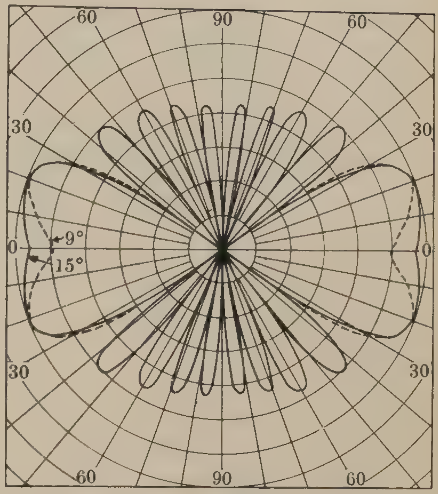


FIG. 11-75 Horizontal directive pattern for five-wavelength aerial at vertical angles of 9° and 15°.

a half-wave aerial except that in adjacent sections of a long-wire aerial the currents must be out of phase. This requirement is met by feeding the long-wire aerial at one end or at a current loop. Tuned feeders are most practical for feeding a multi-band long-wire aerial. Feeding the end of the aerial allows it to operate as a long-wire aerial on all bands but if it is fed at a current loop it only operates as a long-wire aerial on the band for which a current loop exists at the feeding point.

B. Non-Resonant Long Single-Wire Aerials

The resonant long single-wire aerial operates with standing waves of current and voltage along the wire. The non-resonant long single-wire aerial operates with a practically uniform current in all parts of the wire. In its simplest form it consists of a long single wire (Fig. 11-76) grounded at the far end through a resistor *R* having a value equal to the characteristic impedance of the aerial.

The terminating resistor just as in the case of an ordinary transmission line eliminates standing waves. Thus the current decreases uniformly along the wire towards the terminated end. The decrease in current is caused by radiation and resistance loss in the wire. Since the current in such an aerial is relatively small the aerial must be long in terms of the wavelength, to produce good radiation. Greater length of wire is required to radiate the amount of energy that can be radiated from a resonant long-wire aerial which has standing waves. Also the non-resonant long-wire aerial must not be so close to the ground that the return path through the ground will cause cancellation of radiation. If the wire is long it is practically non-resonant over a wide band of operating frequencies.

The terminated non-resonant aerial is practically uni-directional, producing greatest radiation in a direction toward

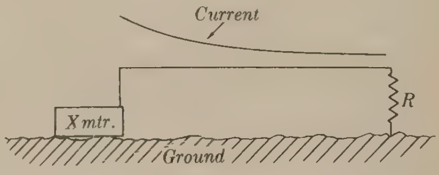


FIG. 11-76 Non-resonant long single-wire aerial.

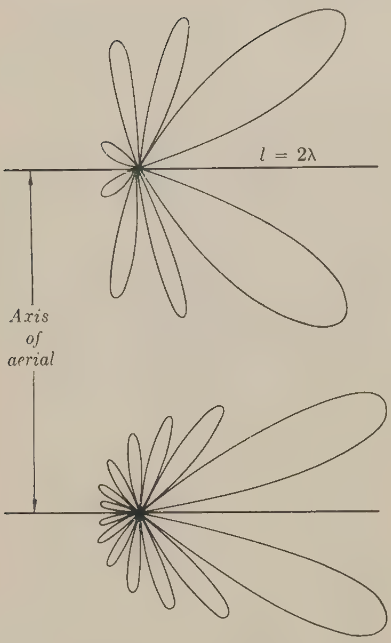


Fig. 11-77 Free-space horizontal directive patterns for two long-wire non-resonant aerials.

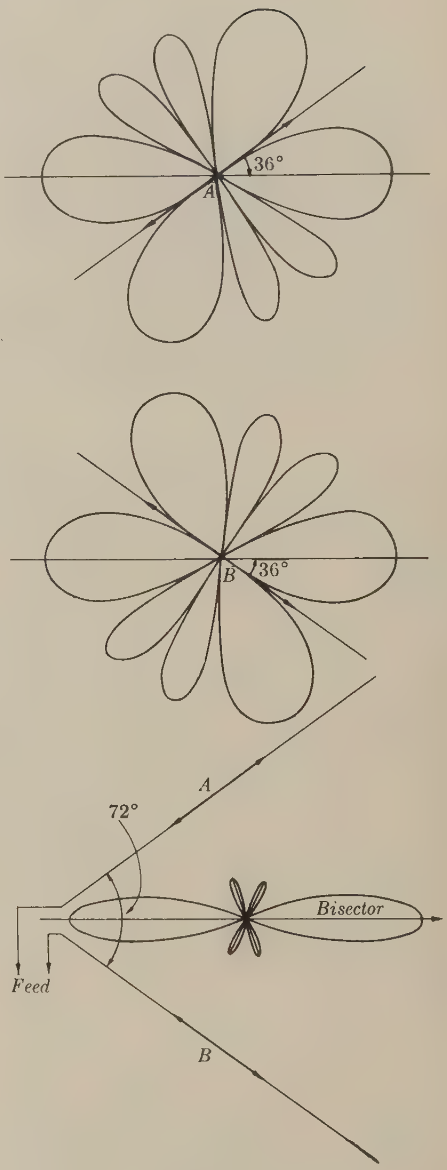


Fig. 11-78 Horizontal *V* aerial.

the terminated end. The directivity varies with the length of the aerial in wavelengths but in every case the radiation in the backward direction is less than in the forward direction. The free space horizontal directive patterns for two typical long-wire non-resonant aerials are shown in Fig. 11-77. The energy which would be radiated in the backward direction from a resonant aerial is absorbed by the terminating resistor of the non-resonant long-wire aerial.

This elementary non-resonant long-wire aerial is not an especially good radiator and for this reason is not often used in this form. However the general principle which provides uni-directional radiation is utilized in the construction of the rhombic aerial (Sec. 11.14) which is a highly practical uni-directional aerial system.

11.13 The horizontal V aerial. In section 11.10 it is shown that the major lobe of radiation from a two-wavelength long-wire aerial occurs in a direction which is 36° to the line of the aerial (Fig. 11-71). If two such aerials are erected to form a V (Fig. 11-78) with a contained angle of 72 degrees, that is, twice 36 degrees, a very effective bi-directional aerial results. If the two long-wire aerials which form the V are excited 180° out of phase by connecting the two-wire feeder to the lines at the vertex of the V, their lobes reinforce along the line bisecting the V angle and tend to cancel in other directions (Fig. 11-78).

The V aerial is simple to build and operate but requires considerable space. It operates satisfactorily over a band of frequencies although it produces optimum results for only one frequency. The gain of the V can be increased by stacking a second V above the first a half wavelength away and feeding so that the legs on one side operate in phase with each other but out of phase with the legs of the other side (Fig. 11-79).

The V aerial is made uni-directional by using a driven reflector consisting of a second V aerial located an odd number of quarter wavelengths behind the first aerial. This driven aerial is excited 90° out of phase to the first aerial. The V aerial can also be made uni-directional by terminating the open ends of the long wires from the V to ground through resistors although this practice is not recommended and not common.

The height of the long-wire legs of the V above ground should not be less than one half wavelength; one wavelength is common in practice. As the length of the aerial, in wavelengths, increases the radiation becomes more directive and more effective. The maximum angle of radiation in the vertical plane is about 15 degrees if the wire length is of the order of five wavelengths; about 25 degrees if the wire length is of the order of two wavelengths.

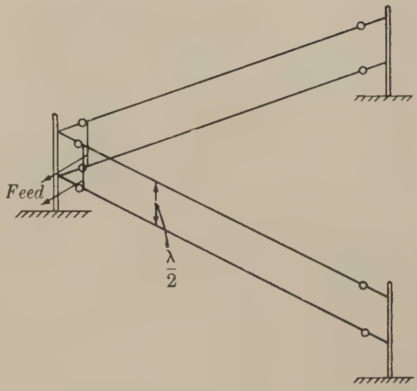


FIG. 11-79 Stacked *V* aerials.

11.14 The rhombic aerial. The *rhombic aerial* (Fig. 11-80) consists of two non-resonant long-wire aerials. The terminating resistor is connected between the far ends of the two wires and is made equal to the characteristic impedance of

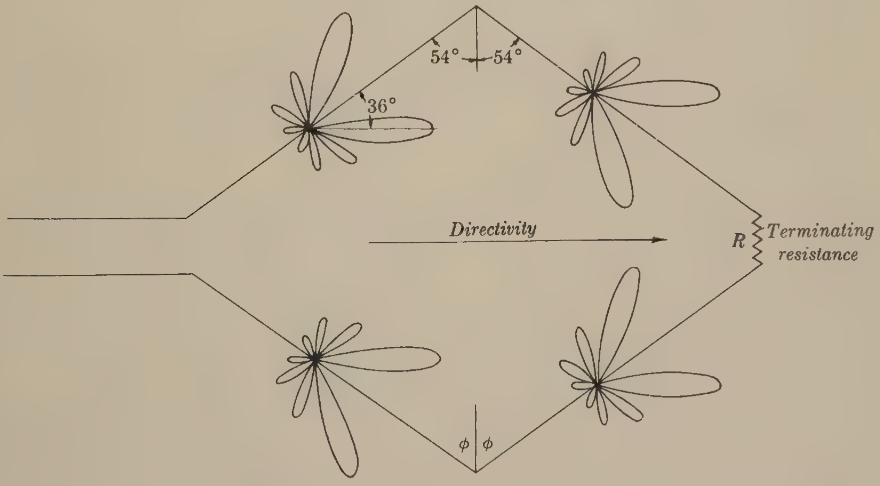
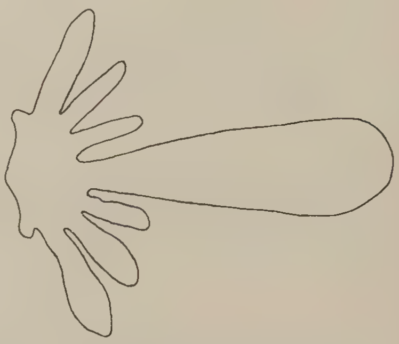


FIG. 11-80 Rhombic aerial.

the aerial as a whole. The rhombic aerial is usually constructed horizontally for economy of pole height and for more satisfactory horizontal polarization at the frequencies for which the aerial is most suitable. The rhombic aerial is a development of the non-resonant long-wire aerial. If each leg of the aerial is made 2 wavelengths long the major lobe of each wire makes an angle of 36° with the wire. The tilt angle is adjusted so that the major lobes of radiation from the wires add in the desired direction. The tilt angle is set at 54° and is the complement of the angle which the main lobe makes with each wire. In directions other than to the right in Fig. 11-80 the radiations from the individual legs tend to cancel. The directional characteristics are not particularly critical with respect to the tilt angle provided the wire lengths are greater than two wavelengths. Typical directive patterns of rhombic aerials are shown in Fig. 11-81. For best results the rhombic aerial should be erected over a soil of good conductivity such as marsh or low lying land.

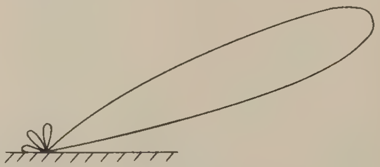
The characteristic impedance of a rhombic aerial is of the order of 700 to 800Ω. The correct value of terminating resistance is about 800Ω and a non-inductive resistor of this value provides operation not far from optimum performance. For convenience the terminating resistance may be placed at the end of an 800Ω line connected to the end of the aerial. This permits placing the terminating resistor at a point which is convenient for adjustment rather than at the top of a pole. The line length is not critical since it is non-resonant.

The rhombic has the great advantage of operation over a wide frequency band without readjustment and requires only low masts. In addition the rhombic may be connected directly to a feeder with a characteristic impedance of 650 to 700Ω.



Horizontal directivity of typical rhombic aerial

(*a*)



Vertical directivity of typical rhombic aerial

(*b*)

Fig. 11-81 Typical directive patterns of rhombic aerials.

11.15 Field strength measurement.

A. Units

The measurement of the strength at a given distance from an aerial provides a means of assessing its efficiency and the directional characteristics of its radiation. The fundamental unit of field strength is the volt per metre but in practice the sub-units, millivolt per metre and microvolt per metre, are more frequently used. The voltages induced in a receiving aerial at a distance from a transmitter are extremely small and for this reason the voltages are usually expressed in microvolts.

The voltage represented in space by vertical electric lines of force is a function of the distance between the horizontal planes at which the measurement is made. Since the field extends upward great distances above the surface of the earth and since wavelengths are usually expressed in metres, it is conventional

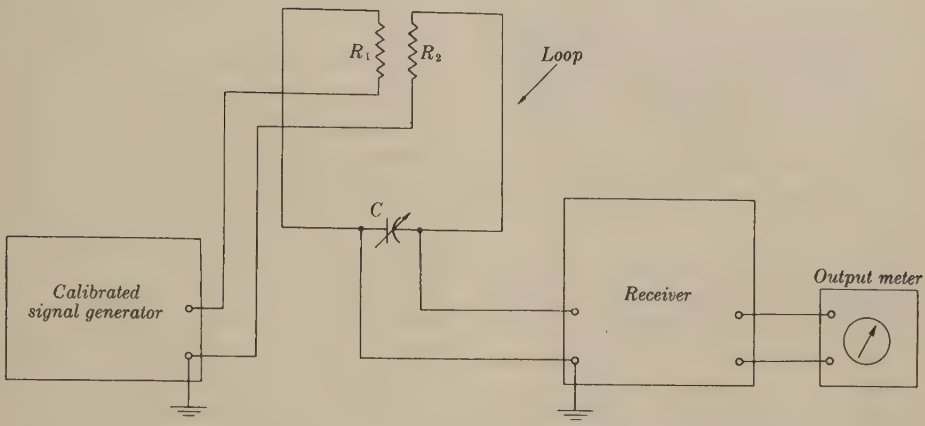


FIG. 11-82 Arrangement of equipment to make a field strength measurement.

practice to express field strength in terms of microvolts per metre. For example, if at a given point a vertically polarized electric field has a strength of 500 microvolts per metre, then a vertical aerial having an effective height of 1 metre erected at that point has a signal voltage of 500 microvolts induced in it. If a vertical aerial with an effective height of 4 metres is erected at that point there is a voltage of 2000 microvolts induced in it.

Field strengths vary over wide limits depending upon ground attenuation. A 500 *KW* broadcast transmitter provides an approximate field strength of 6 volts per metre at a distance of one mile. This is excessive for reception since broadcast receivers provide satisfactory signal output from input signals as low as 2 to 4 microvolts.

B. Method of Making Measurements

The usual method of field strength measurement is one of comparison with a signal of known value. Thus the signal to be measured is received in the usual manner and the reading of the receiver output meter is noted. Then a locally generated signal of the same frequency is introduced into the receiver and the magnitude of the input signal is adjusted to produce the same output on the output meter. In this way the number of microvolts induced in the receiving aerial by the unknown field is obtained. It still remains to find the effective height of the receiving aerial. This problem is solved if an aerial with a known or easily computed effective height is used. In practice a loop aerial finds wide application for field strength measurement below 2000 Kc/s. The effective height of a loop aerial may be calculated from the formula

$$h_{eff} = \frac{2\pi AN \cos\theta}{\lambda}$$

where, h_{eff} = effective height of the aerial in metres
A = area of the loop in square metres
N = number of turns
θ = angle between the plane of the loop and the plane of wave propagation
λ = wavelength of the signal in metres.

Fig. 11-82 shows the arrangement of equipment to make a field strength measurement.

The calibrated signal generator is connected to the centre of the loop through two 100Ω resistors R_1 and R_2 in order to prevent shunting of the signal generator's output by a low impedance which is of the order of 10 ohms when the loop is resonant. Capacitor C is used to tune the loop to resonance at the frequency of the wave the strength of which is to be measured. The loop is connected directly to the receiver. The following procedure is used in taking a field strength measurement.

(i) The loop is rotated until the plane of the loop is parallel to the plane of the wave propagation, a condition which is determined by obtaining the position of maximum signal.

(ii) The signal generator is adjusted to the exact frequency of the signal to be measured and is then turned off.

(iii) The gain control is adjusted to provide a convenient reading on the output meter and this reading is noted.

(iv) The loop is rotated for minimum signal, that is the output meter should read nearly zero, and the signal generator is switched on.

(v) Without disturbing the receiver controls the magnitude of the signal generator is adjusted to provide the same output meter reading as previously noted.

(vi) The value of the signal induced in the loop in microvolts is read from the calibrated signal generator.

(vii) This value is divided by the maximum effective height ($\theta = 0°$), in metres, of the aerial to obtain the field strength in microvolts per metre.

This method of measuring field strength is used at broadcast and lower frequencies and can be employed at higher frequencies. At very high frequencies it becomes difficult to use a loop aerial because of unbalanced effects and distributed capacity in the loop. It is therefore desirable to use a vertical aerial. One method is to use a dipole connected to a concentric transmission line for an aerial and then replace this aerial by a concentric line leading to a signal generator for calibration purposes.

The theoretical radiation patterns of specific aerials have been discussed previously but it must be pointed out that in practice the patterns are somewhat different. The practical radiation directive pattern for a given aerial may be obtained by taking a series of field strength measurements and plotting these readings on polar graph paper to obtain a curve which shows the relative field strength along various lines of radiation.

11.16 Applied mathematics.

A. Length of Half-Wave Aerial

Owing to aerial supports the electrostatic capacity at the ends of a dipole is appreciable. This end effect causes the electrical length to be almost 5% longer than the physical length. The percentage varies slightly with different installations and different frequencies but on an average the length of a half-wave aerial for the range 3-30 Mc/s is given by:

$$\textbf{length (in feet)} = \frac{\mathbf{492 \times 0.95}}{f(\mathbf{Mc/s})}$$

or

$$\textbf{length (in feet)} = \frac{\mathbf{468}}{f(\mathbf{Mc/s})}.$$

B. Length of Reflector

Maximum reflection is obtained from a reflector which is about 5% longer than the half-wave aerial. That is for a reflector:

$$\textbf{reflector length (in feet)} = \frac{\mathbf{492}}{f(\mathbf{Mc/s})}.$$

C. Length of Director

Maximum directivity is obtained from a director about 4% shorter than the half-wave aerial. That is, for a director:

$$\textbf{director length (in feet)} = \frac{\mathbf{450}}{f(\mathbf{Mc/s})}.$$

11.17 Questions and problems.

1. Define
   (a) aerial gain
   (b) aerial directivity.
2. State the directional characteristics of
   (a) a horizontal dipole
   (b) a vertical dipole.
3. Draw the free space polar diagram of radiation for
   (a) a horizontal dipole
   (b) a vertical dipole.
4. What is the main characteristic of a folded dipole?
5. Why are dipoles often constructed with large-diameter elements?
6. State the fundamental principle of a turnstile aerial.
7. What is the main advantage of the conical or cage type dipole?
8. (a) Explain fully, with the aid of diagrams, the principle of the collinear aerial.
   (b) Draw diagrams to illustrate its directivity.
9. (a) Explain fully, with the aid of diagrams, the principle of the broadside aerial.
   (b) Draw diagrams to illustrate its directivity.
10. (a) Explain fully, with the aid of diagrams, the principle of the end-fire aerial.
    (b) Draw diagrams to illustrate its directivity.
11. What is a parasitic element?
12. Define
    (a) reflection
    (b) direction.
13. Explain the principle of operation of
    (a) a reflector
    (b) a director.
14. Draw approximate polar diagrams to illustrate the directivity of
    (a) a Yagi array
    (b) a Lazy-*H* array.
15. State the reflector property of the parabola.
16. Name 4 types of parabolic reflectors and make diagrams to illustrate each.
17. Classify the methods of feed of a parabolic reflector.
18. Why is direct radiation from a dipole which feeds a parabolic reflector eliminated?
19. How does the effect of ground modify the directional performance of an aerial?
20. What is an image aerial?
21. Draw the directional pattern of a Yagi as modified by ground effects.
22. State the principle on which the $\frac{\lambda}{4}$ aerial operates.
23. (a) What is meant by a ground-plane aerial?
    (b) Give typical examples.
24. What is meant by
    (a) a resonant long single-wire aerial
    (b) a non-resonant long single-wire aerial?
25. Draw the directional pattern of a non-resonant long (2λ) single-wire aerial.
26. Draw a diagram of a rhombic aerial and indicate its directivity.
27. What is the unit of field strength measurement?
28. Explain the meaning of a field strength of 600 microvolts per metre.
29. Describe the comparison method of making a field strength measurement.

CHAPTER XII

PROPAGATION

In chapter X it has been shown how radiation takes place from an aerial and how the energy is radiated in the from of an electromagnetic field which travels out into space with the speed of light, 3×10^8 meters per second or 162,000 nautical miles per second. It has also been shown that this electromagnetic field consists of a magnetic field and an electric field, the two being in phase, at right angles to each other and both at right angles to the direction of travel. Within a few wavelengths from the transmitting aerial both the lines of magnetic and electric force are appreciably curved, but at great distances this curvature becomes so slight that the front presented to a receiving aerial becomes almost a plane. This plane is called the *wave front* and is shown in Fig. 12-1 (a). In addition, it has been seen that one of the essential properties of a propagated wave is its *polarization.* The *plane of polarization* is that plane containing the line of propagation and the electric lines through the line of propagation, Fig. 12-1 (a). If the plane of polarization is vertical, that is, perpendicular to the earth's surface, the wave is *vertically polarized.* If the plane is horizontal the wave is *horizontally polarized.* Oblique polarization results when this plane is between the vertical and horizontal.

Vertical polarization does not necessarily mean that the electric lines are vertical. For example, if the line of propagation is slanting down toward the earth and the wave is vertically polarized the plane of polarization is vertical but the electric lines in this plane are not vertical, Fig. 12-1 (b).

As the energy is radiated from a transmitting aerial and propagated, in general, some electromagnetic energy travels out close to the surface of the earth and is called the *ground wave* (or *direct ray*). The rest of the electromagnetic energy is concentrated upward at some angle and this constitutes the *sky wave* (or *indirect*

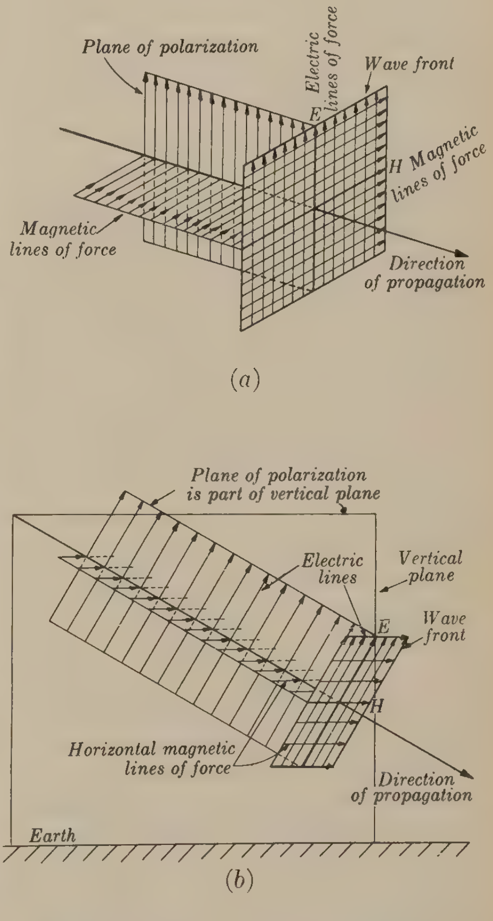


FIG. 12-1 Wave front and plane of polarization of an electromagnetic wave (a) Electric lines vertical (b) Vertically polarized wave in which the electric lines are not vertical.

ray). The sky wave would be lost completely as far as communication is concerned if it were not reflected back to the earth by layers of ions and electrons that exist about 50 to 250 miles above the surface of the earth. To facilitate understanding of ground wave and sky wave propagation, the two waves are discussed separately.

12.1 The ground wave. The ground wave is that part of the radiated energy which is propagated close to the ground. The ground wave follows the irregularities and curvature of the earth's surface in much the same way as electromagnetic waves are guided by transmission lines. The aerial is arranged in practice so that the radiated ground wave is vertically polarized to prevent severe attenuation of the wave. If such a ground wave were obliquely or horizontally polarized the conductivity of the earth would tend to short-circuit any electrostatic component parallel to its surface. Or in terms of the magnetic component of the wave, if the magnetic field were to cut the earth's surface, the circulating currents so formed would dissipate this component.

If the electric lines in the ground wave are almost vertical these effects are not appreciable. The electric lines are never vertical, however, unless the wave is travelling over a surface which has infinite conductivity or is travelling in free space. Since the conductivity of the earth is not infinite and the electric lines are not quite vertical, the ground wave induces voltages and currents in the ground that reduce the energy of the wave. The amount of *attenuation* depends not only upon the conductivity of the earth but also upon the frequency of the radiated wave. It is of considerable importance to know the amount of the attenuation in order to determine the distance the ground wave will travel before the signal becomes too weak to be detected by a receiver.

Soil conductivity depends upon the various types of surface over which transmission is carried out. The conductivity and hence the resistivity varies over wide limits and the following table gives a rough measure of relative resistivity of various surfaces.

SURFACE	RESISTANCE OHMS PER CM. CUBE
Sea water	10
Wet earth	10^2
Fresh water	10^5
Dry earth	10^6
Dry sand	10^8

Thus the distance that a ground wave travels without appreciable attenuation depends partly on the nature of the surface over which it is propagated. A particular wave travels much farther over wet ground or the sea than over dry ground or desert.

In addition, the attenuation of radio waves increases as the frequency increases. At relatively low frequencies of the order of 30 to 300 Kc/s, the ground wave covers long distances, as much as a few thousand miles, with little attenuation. At medium frequencies, 300 to 3000 Kc/s, the wave is more quickly attenuated, the ground wave covering perhaps several hundred miles. High-frequency waves, 3 to 30 Mc/s, are highly attenuated, being completely absorbed after travelling only a few miles. The actual distance covered in each case depends, of course, on the power of the transmitting station. Fig. 12-2 shows the range for a vertically polarized ground wave for frequencies of 30 to 3000 Mc/s over earth or sea for a radiated power of 1 kilowatt.

The range at which ground-wave communication can be maintained over land or sea is important for navigation purposes on medium frequencies. The ground-wave coverage is somewhat less during the day than at night, but

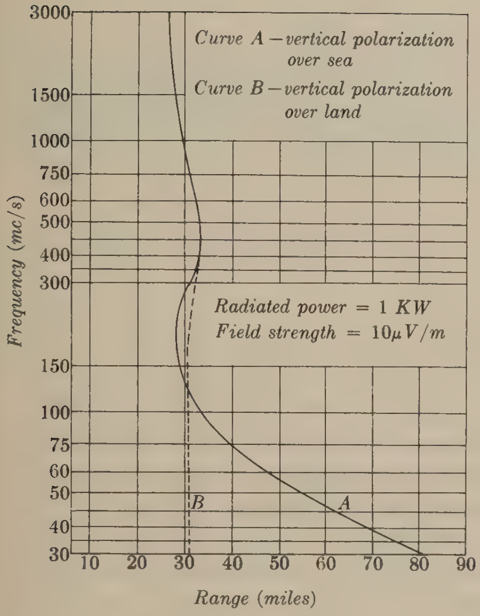


FIG. 12-2. Range of vertically polarized ground wave over earth and sea.

ground-wave reception may be relied upon at any given time. Such services as broadcast stations and transoceanic radio telephony rely mostly on the ground wave and therefore employ the low and medium frequencies.

12.2 The sky wave. The sky wave is that part of the energy propagated upward. The sky wave would be of little use in communication, however, if it continued upward and did not return to earth. Fortunately, this is not the case and the sky wave is reflected back to the earth by the *ionosphere* or *Kennelly-Heaviside* layer. This phenomenon of reflection makes the sky wave of first importance in maintaining world-wide radio communication. To understand how the sky wave is reflected, it is first necessary to understand the nature of the ionosphere.

12.3 The nature of the ionosphere. The ionosphere is an ionized region that exists in the outer portion of the earth's atmosphere. It consists of three main ionized layers designated by the letters E, F_1 and F_2. These layers extend upward from about 50 miles to 250 miles. The lowest or *E layer* is also known as the *Kennelly-Heaviside layer* after two physicists who, independently but simultaneously, predicted the existence of this region.

The air of the lower atmosphere in normal condition is almost a perfect insulator and has negligible effect upon the electromagnetic waves which pass through it. However air, like other gases, under proper conditions may become ionized, that is, electrons may become isolated from their molecules leaving positively charged ions. Ionization may occur as a result of electrons being released from a molecule not only by collisions with other fast moving electrons but also by other types of radiation such as X-rays, ultraviolet rays and cosmic rays. During the daytime the higher atmosphere becomes ionized mainly by the ultraviolet rays of the sun. After ionization the atoms, ions and electrons are still in continual motion producing collisions and an electron may recombine with a positive ion to restore it to a molecule once more. As a result of this recombination process a molecule once ionized does not necessarily remain so. In the lower parts of the atmosphere collisions are so frequent that this recombination prevents appreciable ionization. In addition the ultraviolet rays from the sun are almost entirely absorbed by the upper atmosphere. Consequently little ionization results below a height of 50 miles. From a height of 50 miles to 250 miles conditions are satisfactory for considerable ionization. Above 250 miles the number of air particles is so small that the density of ionized particles is negligible.

The ionosphere consists of several ionized layers due to the fact that the different gases in the earth's atmosphere ionize at different pressures. In addition, ionized layers are to some extent the result of ionizing rays penetrating to different depths. As a result, the heights of these layers above the earth vary from day to day, month to month and from year to year. The lower or *E* layer varies from a height of 50 to 90 miles for its lower boundary. The other layers, F_1 and F_2, only occur separately during daytime and merge into one layer, usually designated by the letter *F*, at night. The *F*, F_1 and F_2 layers are always above the *E* layer. The approximate limits of the heights of the various layers of the ionosphere are given in the following table:

LAYER	LIMITS OF HEIGHT OF LOWER BOUNDARY IN MILES
E	50 to 90
F (at night)	110 to 250
F_1 (day)	90 to 155
F_2 (summer day)	155 to 220
F_2 (winter day)	90 to 185

12.4 Reflection of the sky wave by the ionosphere. The radio waves which make up the sky wave have their paths of travel changed as they pass through the layers of the ionosphere. Often the path of the sky wave is changed enough to make the wave return to earth; sometimes the sky wave penetrates all of the layers and is lost.

The effect of the ionosphere on the sky wave may be seen by a consideration of Fig. 12-3. Suppose a sky wave is radiated from the aerial of transmitter *T* and travels along the path indicated. The path of travel is a straight line until the wave reaches the ionosphere. The wave path now becomes bent and the bending increases as it proceeds into the denser portion of the ionosphere. This bending always takes place away from a region of higher ion density. Thus, the wave path through the ionosphere becomes curved as shown and the wave returns to earth. In practice it is convenient to think of the wave as being reflected and to assume the path to be along the straight lines *TA*, *AR*. This assumption is made in measurements of the heights of the layers of the ionosphere. Thus a signal transmitted from a place *T* on the earth may be received at a point *R*. The distance between *T* and *R* may be far greater than that which could be covered by ground wave and hence communication is made possible by reflection of the sky wave.

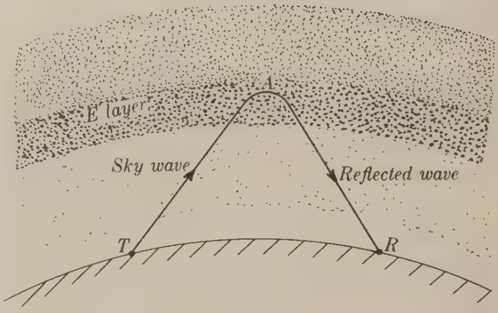


FIG. 12-3 Reflection of the sky wave by the *E* layer.

12.5 Effect of frequency on reflection. Wave reflection from a layer of the ionosphere depends upon the following three factors:

(i) the frequency of the transmitted wave
(ii) the ion density of the layer
(iii) the angle at which the sky wave strikes the ionosphere.

If the frequency of a reflected wave is gradually increased it is found that a *critical frequency* is reached above which the wave is no longer reflected back to the earth. The path travelled by a radio wave of greater frequency than this critical frequency is shown in Fig. 12-4. It is seen that the path of the wave begins to bend, as before, on reaching the ionosphere at *A* but the bending is much less than for a lower frequency. As the frequency of the transmitted wave is increased further the bending is even

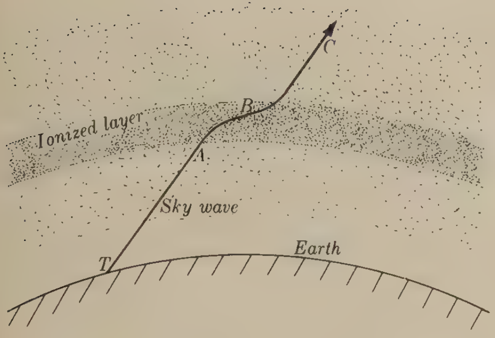


FIG.12-4 Path travelled by a sky wave in penetrating an ionized layer.

less pronounced. The bending is still away from the region of greater ion density as shown. On reaching point B the wave is in the region of greatest ion density and again the bending takes place away from this very dense region. As a result, the wave is bent in a direction away from the earth and travels toward C where it leaves this ionized layer. The wave travels outward and will not return to the earth unless it is reflected from another higher layer.

12.6 Effect of ion density on reflection. In Fig. 12-4 conditions of ion density and frequency are such that the sky wave penetrates the ionosphere and no reflection by this layer occurs. However, if the density of the ionosphere increases for some reason, it may cause sufficient bending of the wave to produce reflection even at this higher frequency and communication at this frequency would be possible. Since the ion density of the ionosphere is a function of the atmospheric conditions it cannot be controlled. As a result, to control reflection it is necessary to control the frequency of transmission.

An important effect of the ion density is the attenuation or absorption of energy from the wave. If the attenuation is great it becomes an important factor in limiting radio transmission. The amount of energy dissipated by a radio wave depends upon the distance it travels through the ionosphere and the density of the ions and the air molecules in the layer. Since ultraviolet rays from the sun cause ionization at a lower level in the daytime than at night, and since there are more air molecules at the lower levels, absorption of energy is greater during the daytime than during the night. The maximum absorption occurs at a frequency of about 1400 Kc/s; the farther the frequency is from this frequency the less the attenuation.

12.7 Effect of angle of incidence on reflection. The angle of incidence is the angle between the line of propagation and the normal to the ionosphere, Fig. 12-5 (a). This is the same as the angle between the line of propagation and the normal to the earth. The effect that the angle of incidence has on the direction of the reflected wave may be seen from a consideration of Fig. 12-5 (b). Sky waves are propagated with different angles of incidence along the lines TA, TB and TC. The sky wave transmitted along direction TA meets conditions which permit it to penetrate the ionosphere.

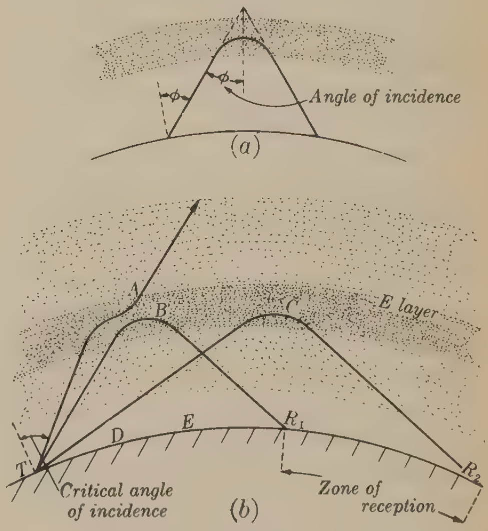


FIG.12-5 Reflection of sky waves (a) Angle of incidence of sky wave at ionosphere (b) Effect of angle of incidence on reflected wave.

However, the sky wave travelling in direction TB, although of the same frequency, strikes the ionosphere at a more glancing angle and is reflected back to the earth. Waves which make greater angles of incidence than that of TB are reflected. The greatest angle of incidence for which the sky wave, for a given frequency, may strike the ionosphere and be reflected back to earth is called the *critical angle of incidence*. Waves that strike the ionosphere at smaller angles of incidence than the critical angle penetrate it; those that strike at larger angles are reflected. The higher the frequency of the wave, the greater the critical angle of incidence and hence the lower the angle of elevation must be to produce reflection.

Thus there is an area from R_1 to R_2 called the *zone of reception* in which it is possible to receive a signal transmitted from T by means of a reflected sky wave. Since the ground wave is completely attenuated before reaching D and since there is no sky wave returned in the area from D to R_1, reception is impossible in this area. This area where reception is not possible is called a *zone of silence* or *skip zone*. The distance TR_1, from the transmitter to the place where the sky wave is first returned is known as the *skip distance*.

When a reflected wave is returned to the earth, it is in turn reflected by the earth and because of its relatively low attenuation it continues to skip around the earth, making world-wide communication possible. This is represented in Fig. 12-6 which shows a two-skip transmission effected by the E layer.

In addition to the above effects, a wave which penetrates the E layer may be reflected by the F layer and return to the earth at a very great distance from its starting point. Such a wave may, as before, be reflected from the earth and produce a multi-hop transmission. Indeed, such a wave may be reflected back and forth between the ionized layers and travel great distances in this way. Again, due to changing conditions over a long path, reflections may occur alternately from the E or the F layers.

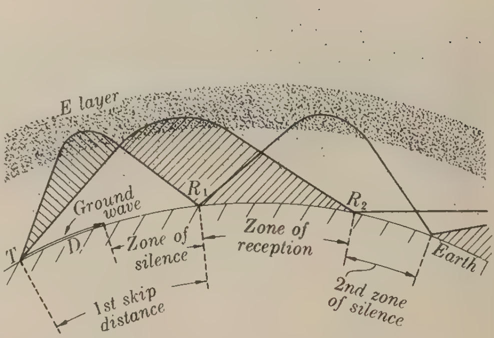


Fig. 12-6 Two-skip transmission effected by E layer.

Of course, if the transmitted wave strikes the layer at too large an angle of incidence the attenuation due to ground losses again becomes appreciable and no reflection results. Thus, there is an upper limit for the value of the angle of incidence as well as a critical value at the lower limit.

Fig. 12-7 shows the angles and terms which are commonly used in discussion of the effect of the ionosphere on a radio wave. The height, h, is called the *virtual height* of the ionosphere since it is not the true height. The following are the designations for the angles indicated

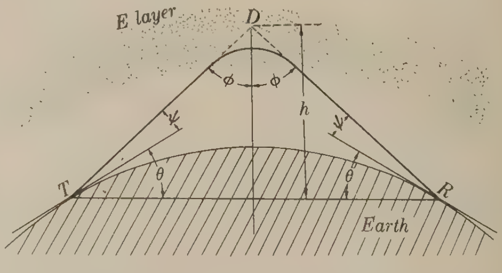


Fig. 12-7 Designation of angles used in discussion of a reflected wave.

in the diagram:

ϕ is the angle of incidence

Ψ is the angle of elevation at the earth's surface

θ is the angle of depression at the earth's surface.

The lower limiting value of Ψ is often taken as lying between 3° and 5°.

12.8 Variations in the ionosphere. It has been stated in section 12.3 that the heights of the layers of the ionosphere may vary from day to day, from month to month and from year to year. These changes are largely a result of the changes in the radiation from the sun. Hence the changes in the ionosphere are for the most part changes in ion density. If the ion density during the day period becomes a maximum at a lower level than at night, the virtual height of the ionosphere in effect is changed. The F_2 layer shows a considerable variation in height while the other layers show only a moderate variation.

In general, the variations in the ionosphere can be predicted quite accurately. The variations in the E layer occur regularly from day to day and depend almost entirely on the position of the sun. During the day ionization density is highest while during the night it is lowest. Again the density is greater during summer than during winter. In addition to these variations the density also varies with sunspot activity and is greatest during the most active periods. Thus allowances in predictions must be made for changes resulting from the 11-year solar cycle. The approximate variation in the vertical heights of the E, F_1 and F_2 layers during the day is shown in Fig. 12-8.

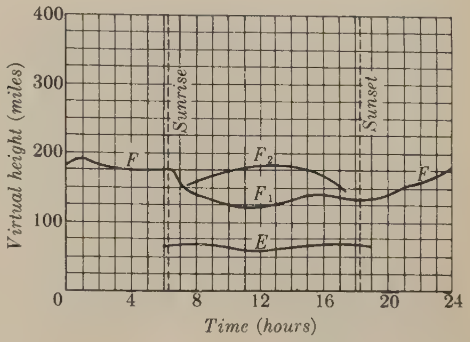


FIG. 12-8 Hourly variation in virtual height of E, F_1 and F_2 layers.

12.9 Maximum usable frequency. The highest frequency for which waves sent vertically upward are returned to the earth is called the *critical frequency*. However, it has been seen that waves of higher frequencies than the critical frequency may be reflected if radiated at an angle of incidence greater than zero. If a wave meets the ionosphere with an angle of incidence ϕ (Fig. 12-7), it has been found that the maximum frequency that is reflected is given by the formula

$$f_{max} = f_o \sec \phi$$

where f_o = critical frequency for vertical incidence.

This maximum frequency which can be used is called the *maximum usable frequency*.

Measurements of critical frequency are taken at many places throughout the world and the values are published in peacetime. This information, together with additional information on layer density, time of day, month of year and year of solar cycle makes it possible to plot graphs which give the maximum usable frequency which can be used to transmit to any distance. Such curves are usually provided for some months in advance. Fig. 12-9 shows one form of such a graph. This graph gives the maximum usable frequency during a period of minimum sunspot activity. Several curves are given for different distances between transmitter and receiver.

12.10 Fading. *Fading* is the name given to variations in the intensity of a received signal and causes variation in the a-f output of a receiver.

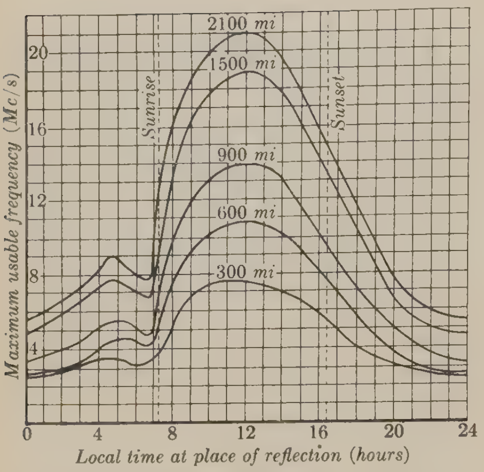


FIG.12-9 Form of graph which shows maximum usable frequencies.

The cause of fading can be understood by a consideration of Fig. 12-10. Signals transmitted at T are reflected from the E and F layers as shown. The received signal at R_1 is made up of a component from the ground wave and a component from the reflected wave. In general, the two components travel different distances from transmitter to receiver. If these signals are in phase they add to produce a loud signal; if they are out of phase they subtract to produce a weak signal. The in-phase or out-of-phase condition depends upon the difference in distance travelled by the ground and reflected waves. The distance travelled by the reflected wave in turn depends upon the height of the F layer. Small changes in this height produce small changes in the length of the path travelled by the reflected wave. These variations in path difference are sufficient to change from the in-phase to the out-of-phase condition and produce corresponding changes in signal strength.

Fading can also occur as a result of interference between two sky waves as shown at R_2. The change in either or both of the layers can produce fading at this receiving station.

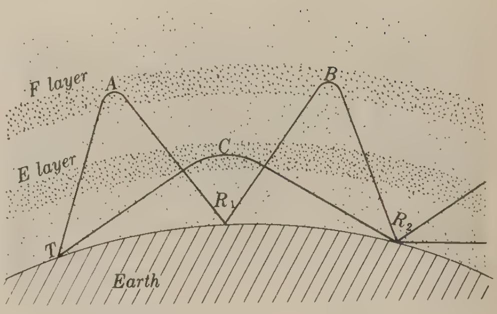


FIG.12-10 Cause of fading.

In addition to the fading described above, some of the sideband frequency components of a modulated signal may fade more than others and cause distortion of the audio output of the receiver. This is known as *selective fading.*

Still another type of fading is sometimes produced by sudden variations in ion density of the ionosphere. The most sudden fading is of this type and it is called *radio fadeout.* Increase in ion density due to radiation from the sun causes a corresponding increase in absorption of the sky wave. The fadeout usually occurs for all frequencies above 1500 Kc/s and may last from several minutes to hours.

Thus fading in a received signal may take place slowly or very suddenly. However, the variations in signal strength caused by the regular variations in the ionosphere discussed previously are not usually considered fading.

12.11 Propagation characteristics in the various frequency bands. The communications frequencies and higher radio frequencies are classified in the table at the top of page 12:9. The propagation characteristics within these bands vary considerably and each is discussed in some detail.

A. Low Frequencies
(*Below* 300 *Kc/s*)

Transmissions in the very-low-frequency range below 30 Kc/s are propa-

Classification of Frequencies

GENERAL DESIGNATION	ABBREV.	FREQUENCY RANGE	WAVELENGTH	WAVE DESIGNATION
Very low frequencies	v-l-f	Below 30 Kc/s	Above 10,000m	Very long waves
Low frequencies	l-f	30-300 Kc/s	10,000-1000m	Long waves
Medium frequencies	m-f	300-3000 Kc/s	1000-100m	Medium waves
High frequencies	h-f	3-30 Mc/s	100-10m	Short waves
Very high frequencies	v-h-f	30-300 Mc/s	10-1m	Metre waves
Ultra high frequencies	u-h-f	Above300Mc/s	1m-10cm	Decimetre waves
			10cm-1cm	Centimetre waves
			1cm-1mm	Millimetre waves

gated to great distances with the *E* layer acting as an almost perfect reflector. Radio waves at these frequencies are reflected back and forth between this layer and the ground and often encircle the earth. Attenuation is relatively low at all times of day and throughout the year. As the frequency increases the penetration into the layer increases and of course the attenuation increases.

B. Medium Frequencies
(300-3000 *Kc/s*)

As the frequency of the transmission is increased from 300 Kc/s there comes a band of frequencies ranging from approximately 800 Kc/s to 2000 Kc/s for which absorption by the ionosphere is greatest. Maximum absorption occurs at a frequency of about 1400 Kc/s. In this band propagation takes place over distances of only a few hundred miles. Attenuation is lower at night and this extends the range but fading is pronounced due to interaction between ground and reflected waves. In addition attenuation is less in winter making increased ranges possible during this season.

C. High Frequencies
(3000 *Kc/s*-30 *Mc/s*)

As the frequency used for transmission enters the high-frequency band propagation over any considerable distance depends entirely upon reflection from the *F* layer. The waves at these frequencies escape through the *E* layer with some attenuation but with little reflection. The attenuation of the *E* layer decreases as the frequency increases above 1400 Kc/s. On reaching the *F* layer the skywaves are reflected provided the angle of incidence is not too small. Since the density of ionization of the *F* layer is different during the day from that during the night, it is important to consider the requirements of daytime propagation and night-time propagation separately.

1. DAY AND NIGHT FREQUENCIES.

The lower frequencies of this band, from 3 Mc/s to 10 Mc/s approximately, are more readily reflected and penetrate the layer less.

The higher frequencies of this band, from 11 Mc/s to 30 Mc/s, penetrate the layer to a much greater depth before reflection takes place. The reflection of waves at these frequencies is less pronounced.

In general, since the density of the ionosphere is greater during the day than during the night, the frequencies from 11 Mc/s to 30 Mc/s are usually useful only during the day when the density is a maximum. These frequencies are known as the *day frequencies.* Since the frequencies from 3 Mc/s to 10 Mc/s are more readily reflected they are used mainly during night transmission and as a result are called *night frequencies.*

2. Daytime Propagation.

During the daytime the E layer has a relatively high ion density and the layer boundaries are fairly well defined. If transmissions at the lower night frequencies are radiated the radio waves penetrate the E layer and are reflected by the F_1 or F_2 layers. However, owing to the high ion density of the E layer, attenuation is appreciable and transmission is only practicable over a range of a few hundred miles.

Transmissions at the higher day frequencies also penetrate the E layer but the attenuation is not as severe. They are reflected from the F_1 layer only after deep penetration. Since the density of this layer is considerably less than that of the E layer the resulting attenuation is still not appreciable. As a result communication over great distances is possible using day frequencies.

3. Night-Time Propagation.

At night the density of the E layer is not nearly as great. The F_1 and F_2 layers recombine to form the F layer which also has a very much reduced density. If the day frequencies are used the sky waves penetrate both the E and F layers and are completely lost. However, the density of the F layer is still sufficiently great to produce reflection of waves transmitted at the night frequencies. Attenuation of these waves is extremely small and world-wide communication is possible.

4. Day and Night Frequencies for 24-Hour Communication.

Since the effective heights of the E and F layers are different for day and night, a ship which is within the first skip by day may be in a zone of silence by night even if the wave is reflected by the F layer (Fig. 12-11). This results from the fact that for a sky wave of given frequency there is a longer skip distance by night, TR_2, than by day, TR_1. Thus to maintain 24-hour communication between a ship and a shore station it is necessary to use two frequencies, the higher one during the day. For example, two related frequencies might be 10 Mc/s and 12 Mc/s. The increase in skip distance resulting from an increase in the height of the F layer at night is compensated for by use of the lower frequency. In practice, it is often necessary to use more than two frequencies to maintain 24-hour service between two places.

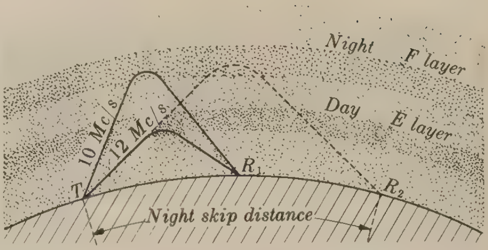


Fig. 12-11 Variation in skip distance with frequency.

D. Very High Frequencies
(Above 30 Mc/s)

For the very high frequencies the sky wave is no longer reflected back to the earth except on rare occasions. In general transmissions of very high frequencies behave more like light waves than like any of the lower frequencies. Thus at the very high frequencies communication depends upon straight line or line-of-sight propagation from the transmitter to the receiver.

However, with the transmitting and receiving aerials located over level ground on a perfectly level plane or on ships at sea, as shown in Fig. 12-12, there is also a reflected path TAR as well as the line-of-sight path. The wave reflected by the sea or earth combines with the direct wave at the receiver. These waves reinforce if they arrive in phase and tend to cancel each other if they arrive out of phase. Since the wave reflected from the sea results from re-radiation caused

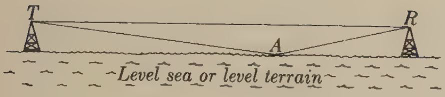


FIG.12-12 V-h-f propagation paths between transmitter and receiver located on level ground.

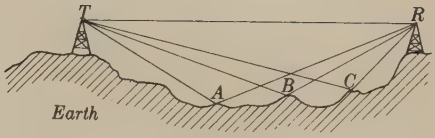


FIG.12-13 V-h-f propagation over uneven terrain.

by currents induced in it by the incident wave, there is a phase reversal at the point of reflection. That is, the reflected wave is 180° out of phase with the direct wave. Since the lengths of the paths of the line-of-sight ray and the reflected ray are almost equal, the rays arrive at the receiver almost 180° out of phase and as a result there is a great reduction in signal.

If the path travelled by the reflected wave is made one half wavelength longer than that of the direct ray, the two arrive in phase at the receiver. If the transmitter and receiver are elevated a considerable distance the difference in the line-of-sight and reflected paths increases. Thus, the two waves arrive at the receiver more nearly in phase with a resulting increase in signal strength. If the elevation is made sufficiently great so that the paths differ by any odd number of half wavelengths the two waves arrive in phase producing a further increase in field strength.

Cancellation or reinforcement of the signal at the receiver depends upon the height of the aerials and their distance apart. If they are located on land the sites should be selected so that the path difference is an odd number of half wavelengths.

The conditions encountered in very-high-frequency propagation over land are generally more complicated than those illustrated in Fig. 12-12 as a result of uneven ground between transmitter and receiver. The line-of-sight wave combines with many reflected waves as shown in Fig. 12-13. Due to these reflections the signal may be considerably reinforced or almost completely cancelled.

As the distance between transmitter and receiver increases the effect of reflected waves becomes less marked. The transmitted wave (Fig. 12-14) is propagated in space between *T* and *A* provided *T* is well elevated. This portion of the path is relatively free of ground losses. Between points *A* and *B* the wave is bent around the earth's surface and is subject to ground (or sea) losses. From *B* to *R* the wave is again propagated tangentially to the earth and losses are small. The signal reaches *R* which is beyond the line-of-sight range as a result of diffraction and refraction. Diffraction is the phenomenon by which sound waves, light waves or radio waves bend around obstacles in their path. The amount of this bending depends upon the size of the obstacle compared to the wavelength of the wave. If the obstacle is small compared with the wavelength, diffraction is pronounced. Thus light waves are diffracted very little since all obstacles are large compared with the wavelength of light. However, many obstacles are small compared with the wavelength of radio waves and bending may occur. At low radio frequencies diffraction is appreciable. At very high frequencies the bending is less but it is still sufficient to enable signals to be received at distances of several miles beyond the line of sight.

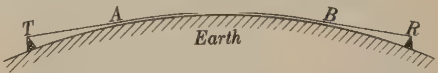


FIG.12-14 V-h-f propagation beyond the line-of-sight path.

Part of the bending of a wave which permits reception at the lower range of very high frequencies some distance beyond the optical range is caused by refraction of the wave. Refraction occurs as a result of changes in the density of air with height and changes in temperature, pressure and amount of water vapour. This refraction in the lower atmosphere is somewhat similar to the bending due to the ionosphere in the upper atmosphere.

As the atmospheric effects which cause refractions are difficult to predict the effects of refraction are equally unpredictable. In general, summer is the best season for increased ranges due to refraction.

In the microwave region, since the wavelength is very short, diffraction and refraction are negligible. As a result only the direct ray can be received and communication is not possible beyond the optical range. At these frequencies directive aerials are usually employed to concentrate the transmitted energy into a very narrow beam in order to produce an increase in signal strength at the receiver.

12.12 Scattered signals. It has been seen that a zone of silence occurs between the farthest point reached by the ground wave and the point where the sky wave is first returned to the earth. However, some signals do reach a receiver located in this zone. These signals are scattered signals which are re-radiated from centres of disturbances in the layers of the ionosphere. The strength of such signals is most uncertain; they may occur at one time but not at another. Fig. 12-15 shows two such scatter sources. In general, there are long-distance scatter centres such as S_1 and near scatter centres such as S_2, both occurring in the E layer. A receiver located at R_1 receives ground-wave signals and scattered signals. However, since the ground wave is strong and the scattered signals are very weak the former drowns the latter and the effect of the scattered signals is seldom noticed. At R_2 in the zone of silence the reception is normally negligible and any transmission received in this region is the result of scattered signals. Scattered signals begin to appear at frequencies of 2500 Kc/s and little scattering is noticeable at frequencies above 10 Mc/s.

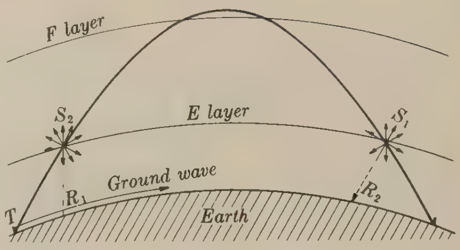


FIG. 12-15 Scatter sources.

12.13 Summary of radio wave propagation characteristics. The table on page 12:13 provides a summary of radio wave propagation for each band of frequencies.

12.14 Questions and problems.

1. Explain briefly how radiation takes place from an aerial.
2. What determines the polarization of a wave?
3. Define
   (a) ground wave
   (b) sky wave.
4. State and explain the kind of polarization used for a ground wave.
5. What effect does the variation in soil conductivity have on the ground wave?
6. What effect does frequency have on the ground wave?

Radio Wave Propagation Characteristics

FREQUENCY RANGE	SKY WAVE	GROUND WAVE
Below 30 Kc/s	The ionosphere acts as an almost perfect reflector and radio waves encircle the earth.	Ground wave travels distances of 1000 miles or more.
30—300 Kc/s	Attenuation increases but is still relatively low. Sky waves make communication possible at distances of from 500 to 8000 miles.	
300—3000 Kc/s	Attenuation increases to a maximum at about 1400 Kc/s. Little sky-wave communication takes place during the day but reception is possible at night at distances from 100 to 3000 miles.	Ground wave is reliable to 50 miles on higher frequencies and 200 miles on lower frequencies.
3000 Kc/s—30 Mc/s	Reflection from *F* layer with little attenuation makes long-distance communication possible with different frequencies for day and night operation. Reception is possible over distances as great as 10,000 miles.	Ground waves are rapidly attenuated. Communication is possible at distances of the order of 15 to 25 miles.
Above 30 Mc/s	There is no reliable sky-wave reflection, no ground-wave propagation; only line-of-sight communication possible making use of direct ray.	

7. What causes the ionization of the earth's atmosphere?

8. Why is there a low density of ionization in the atmosphere below 30 miles and above 250 miles from the earth's surface?

9. What ionized layers exist
   (a) during the day
   (b) during the night?

10. (a) Upon what factors does reflection of the sky wave depend?
    (b) Explain the dependence in each case.

11. Explain the causes of the variations in the layers of the ionosphere from
    (a) day to day
    (b) month to month
    (c) year to year.

12. What factors determine whether or not a radio wave will penetrate the ionosphere?

13. What effect does ion density have on the energy in a radio wave?

14. Explain the effect of the angle of incidence on reflection.

15. Draw a diagram to illustrate
    (a) skip distance
    (b) zone of silence
    (c) zone of reception
    (d) angle of elevation
    (e) angle of depression.

16. Define
    (a) critical frequency
    (b) virtual height
    (c) maximum usable frequency
    (d) critical angle.

17. Explain the cause and effect of the phenomenon of fading.

18. Explain the meaning and cause of
    (a) selective fading
    (b) radio fade-out.

19. (a) What are day and night frequencies?
    (b) Explain why they are made necessary.

20. Classify the radio frequencies into their respective bands and explain briefly the propagation characteristics in each.

21. (a) What are scattered signals?
    (b) Explain what causes these signals.

22. Explain what is meant by
    (a) diffraction
    (b) refraction of radio waves.

CHAPTER XIII

FREQUENCY, PHASE, PULSE MODULATION

13.1 Introduction. In Chapter VI, one system of modulation called amplitude modulation (a-m) is discussed. With this type of modulation the amplitude of the r-f carrier wave is varied in accordance with the audio signal which represents the intelligence to be transmitted. At the a-m receiver the r-f carrier wave is demodulated so that the earphones or speaker reproduces the audio signal originally produced at the transmitting end of the communication system. One of the great disadvantages of this system lies in the fact that both natural and man-made electrical disturbances (called *static*) are received by the receiver at the same time as the r-f carrier wave. These electrical disturbances are r-f waves having frequencies which are similar to those of the transmitter signals and having amplitudes which vary greatly. Thus the demodulation process results in both the audio signal and noise being heard in the phones or speaker. To eliminate this interference, a system of modulation is necessary in which some characteristic of the r-f carrier wave other than its amplitude is varied in accordance with the audio signal.

An r-f carrier wave has two other characteristics besides amplitude, namely, frequency and phase. The system of modulation in which the frequency of the r-f carrier wave is varied in accordance with the audio signal is called *frequency modulation* (*f-m*). Any amplitude variations caused by static can be removed before demodulation without affecting the desired modulation. In this way frequency modulation overcomes the defect which is inherent in amplitude modulation.

The transmitters and receivers required to produce and detect an f-m radio wave are not completely different from those required for an a-m radio wave. The differences, which lie in the modulation stages of the transmitter and the detection stages of the receiver, are discussed in later sections. In the next section a simple f-m transmitter is considered in order to present a clear picture of the character of an f-m wave.

13.2 A simple f-m transmitter. The frequency of an oscillatory circuit depends on the values of its inductance and its capacitance. Hence the frequency of such a circuit may be varied by arranging that this inductance or capacitance be varied. A simple method of allowing voice frequencies to vary the oscillator capacitance is illustrated in Fig. 13-1. The circuit is a simple parallel-feed Hartley oscillator with a capacitor microphone M connected in parallel with its tuning capacitor C_1. This type of microphone is electrically a variable capacitor. The diaphragm D, Fig. 13-2 (a), is one plate of the capacitor and is made to vibrate at an audio rate by allowing

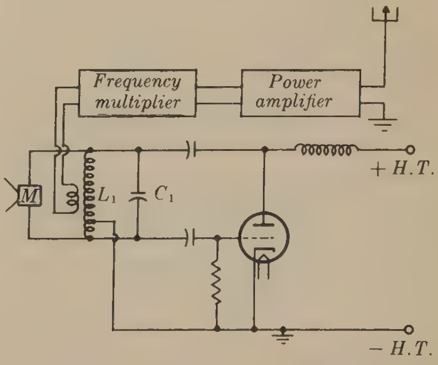


FIG. 13-1 Simple f-m oscillator.

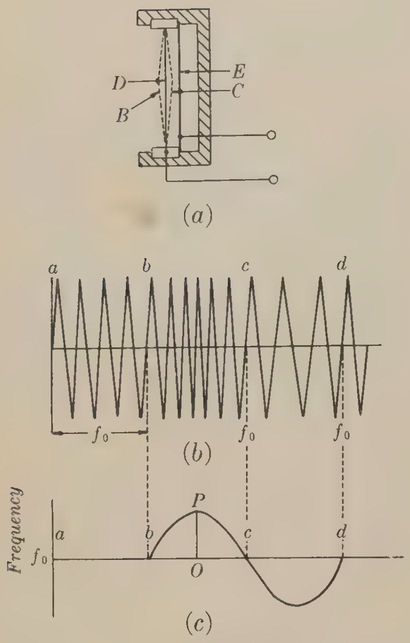


FIG. 13-2 Frequency modulation by means of a capacitor microphone.

sound waves to strike it. The other plate E of the capacitor remains stationary. Since the capacitance depends, in part, on the spacing between D and E, the vibration of D varies the capacitance and hence the frequency of the oscillator. A more detailed understanding of the operation of the circuit is provided by the diagrams of Fig. 13-2 (b) and (c).

When no sound waves strike the microphone, its capacitance remains constant. The frequency of the oscillator remains fixed at a value f_0 determined by the inductance of L_1 and effective capacitance of M and C_1 in parallel. As long as this condition exists, the r-f carrier is unmodulated as shown in the region ab in Fig. 13-2 (b) and (c). When a rarefaction enters the microphone, the diaphragm D is forced to move outward, away from E. The capacitance of M decreases and so the frequency of the oscillator rises above the frequency f_0. This is shown in the region bc of the figure. When a compression enters the microphone, the diaphragm is forced toward E and the capacitance of M increases. This causes the frequency of the oscillator to fall below f_0 as shown in the region cd of both graphs. Hence each time the diaphragm moves from its initial position, out to B, in to C and back to its initial position, the resonant frequency of the oscillator undergoes one cycle of change bcd. In other words, the oscillator frequency is given one cycle of variation for each complete cycle of the sound wave. If the sound wave has a frequency of 1000 c/s then the oscillator frequency undergoes 1000 cycles of change each second. The amount by which the frequency changes from its value f_0, which exists when there is no modulation, is indicated by the amplitude OP in Fig. 13-2 (c) and is called the *frequency deviation* (f_d). The frequency of the f-m transmitter without any modulating input is referred to as the *resting, centre, mean,* or *carrier frequency.*

The amount of deviation produced by this system of modulation, or by any present system of modulation is insufficient for proper detection by an f-m receiver. The amount of deviation must be increased by using one or more frequency multiplier stages (Fig. 13-1) between the oscillator and the power amplifier. An example will serve to illustrate this point. Suppose the output of the oscillator varies in frequency from 5.99 Mc/s to 6.01 Mc/s. The deviation is then .01 Mc/s. If the frequency multiplier is a tripler, its output varies from 17.97 Mc/s to 18.03 Mc/s about an unmodulated frequency of 18.00 Mc/s. The deviation has therefore been tripled from .01 Mc/s to .03 Mc/s. At present, the f-m broadcast band extends from 88 Mc/s to 108 Mc/s. Each channel in the band extends over 200 Kc/s and the maximum deviation for modulation is 75 Kc/s above and below the unmodulated frequency.

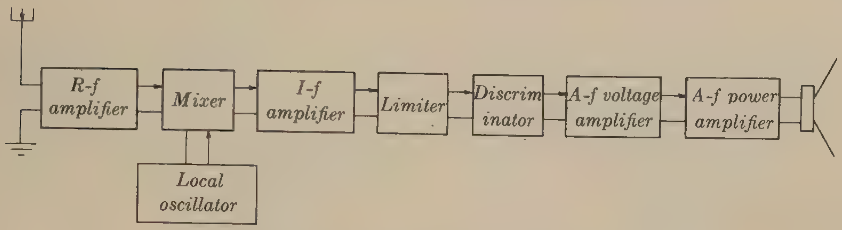


FIG. 13-3 Block diagram of f-m receiver.

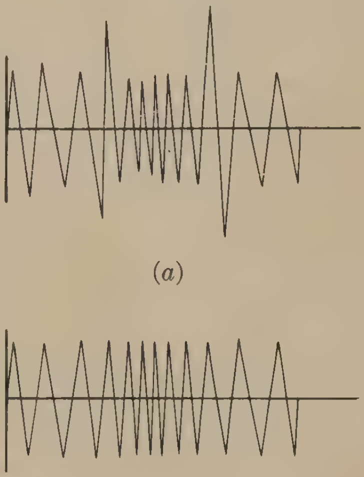
13.3 The f-m receiver. It is possible to use a regenerative, t-r-f or superheterodyne type of receiver for f-m reception. However, the superheterodyne type is used almost exclusively because of the high gain required to amplify the input signal to the level required for proper operation of the demodulating circuits. A block diagram of an f-m receiver is shown in Fig. 13-3. It is similar to the block diagram of an a-m receiver with two exceptions. The second i-f amplifier is replaced by a *limiter* in the f-m receiver. This stage provides an output signal of constant amplitude to the following stage. The second detector is called a *discriminator* in the f-m receiver. The output of the discriminator is the a-f signal. Consequently the a-f circuits may be identical for f-m and a-m reception.

While the r-f, l.o. and i-f stages of the f-m receiver are similar to those in an a-m receiver, they differ somewhat in circuit design and construction because of the higher frequencies of the signals which are amplified. The r-f tuned circuits are physically smaller and both the r-f and i-f circuits are more broadly tuned to provide a bandwidth of about 150 Kc/s.

The limiter and discriminator circuits are described in the following sections.

13.4 The limiter. Although the transmitter signal leaves the transmitter with constant amplitude, the received signal may vary considerably in amplitude due to electrical interference and fading. The undesirable fluctuations in signal strength are amplified by the r-f and i-f stages along with the signal, Fig. 13-4 (a). The desired signal of varying frequency and constant amplitude is shown in Fig. 13-4 (b). The purpose of the limiter is to convert the signals of varying amplitude to signals of constant amplitude at the input to the discriminator.

A common type of limiter circuit is shown in Fig. 13-5. The valve is an r-f pentode operated with an anode potential much lower than the normal value. The cathode is connected to ground and the only bias is that provided by the grid-leak capacitor C_4 and resistor R_1. R_2 and C_3 are the screen voltage dropping resistor and by-pass capacitor. The



(b)

FIG. 13-4 Voltage waveforms of i-f signal (a) with static present (b) without static.

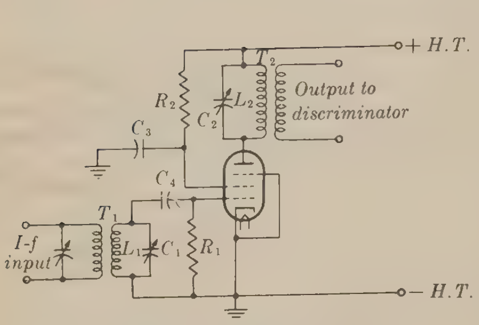


FIG. 13-5 Limiter circuit.

i-f signal is transformer coupled to the tuned input circuit L_1C_1. The output signal is developed across the primary circuit L_2C_2 of the tuned transformer T_2 and transformer coupled to the following stage.

The circuit operates on the principle of anode current saturation. Since the anode voltage is low, only a small positive grid potential is required to cause saturation of the anode current. Any further increase in grid potential is ineffective in changing the anode current. When the signal drives the grid negative, the anode current decreases until cut-off is reached. At this point any further increase in negative grid potential is ineffective in changing the anode current. As a result, the anode current is limited between cut-off and saturation values and the output voltage is correspondingly limited in amplitude (Fig. 13-6).

The time constant of R_1C_4 determines the characteristics of the limiter with respect to slow and rapid variations in amplitude. A short time constant is most effective in limiting noise pulses as they are usually rapid variations. A long time constant is desirable for smoothing out more gradual fluctuations in amplitude. As a result, a compromise time constant of about 10 microseconds is selected for R_1C_4 in most limiters.

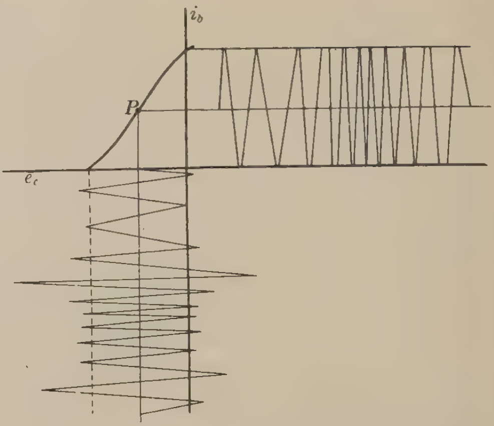


FIG. 13-6 e_{c}-i_b characteristic and operation of limiter.

The limiter is most effective if preceded by several stages of amplification to ensure that the signal amplitude is large enough to cause the limiting action. The limiting effect can be improved by the use of two limiters in series. A constant output over a wide range of input signal amplitudes is obtained from the second limiter because the voltage at the input to this stage is already partially limited. In addition, a grid circuit with short time constant may be used in one stage to limit rapid fluctuations and a circuit with long time constant in the other circuit to limit the effect of slow variations in signal amplitude.

13.5 The discriminator. The *discriminator* in an f-m receiver performs the same function as the second detector in an a-m receiver. The discriminator converts the frequency-modulated i-f signal to an amplitude-modulated signal. This latter signal is then detected to provide an a-f output like that from an a-m detector. After amplification, the a-f signal is made to operate the phones or speaker which reproduces the sound waves falling upon the microphone at the transmitting station.

A. A Simple Discriminator Circuit

Fig. 13-7 (a) shows a simplified discriminator circuit. The tuned circuit 1 is broadly tuned to the i-f carrier frequency f_0 while the tuned circuit 2 is

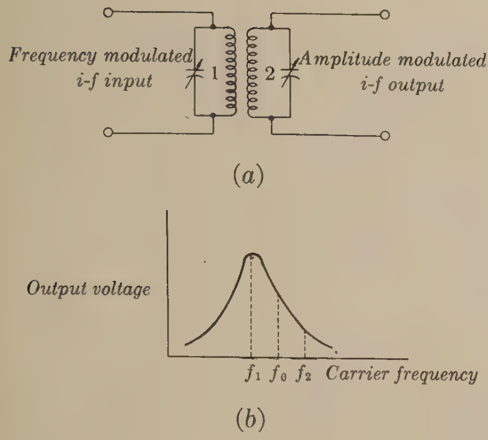


FIG.13-7 Simple discriminator (a) Circuit diagram (b) Response curve of secondary circuit.

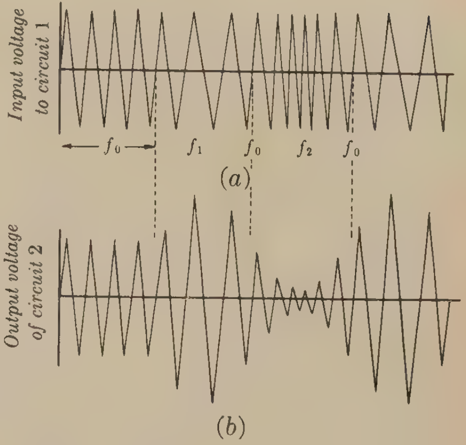


FIG.13-8 Voltage waveforms of simple discriminator (a) Frequency-modulated i-f input (b) Amplitude-modulated i-f output.

sharply tuned to a slightly lower frequency f_1. The response curve of the circuit 2 is shown in Fig. 13-7 (b) and the waveforms of input and output voltages are shown in Fig. 13-8. As the frequency of the i-f input voltage decreases below the mean frequency f_0, the secondary circuit 2 becomes more closely tuned to resonance. Hence the response of circuit 2 increases and the amplitude of the i-f voltage developed across circuit 2 increases. When the frequency of the i-f input rises above f_0, circuit 2 is tuned farther from resonance. Hence in this case the response of circuit 2 falls off and the amplitude of the i-f voltage developed across circuit 2 decreases. Thus as the frequency of the i-f voltage varies from f_1 to f_2 the response of the circuit and hence the output i-f voltage amplitude varies in a corresponding way as shown. One cycle of change in the frequency of the i-f input voltage produces one cycle of change in the amplitude of the i-f output voltage. In other words the frequency of the amplitude variations in the i-f output is equal to the rate of frequency variation in the i-f input voltage, which is also equal to the frequency of the modulating a-f at the transmitter. The greater the frequency deviation (f_2-f_0) the greater the change in response and hence in the variation in amplitude of the output voltage.

In order to prevent distortion from being introduced by the discriminator, the i-f signal in circuit 1 must remain constant in amplitude. This requires that the circuit present the same load impedance to the valve over the range of frequencies from f_1 to f_2. Circuit 1 must therefore be very broadly tuned to provide a flat response over a range of 75 Kc/s on each side of f_0.

B. A Simplified Detuned-Circuit Discriminator

A practical discriminator circuit which performs the complete demodulation process of the i-f signal is shown in Fig. 13-9. It consists of an i-f transformer T_1 with a centre-tapped secondary winding. The two halves of the secondary are L_1 and L_2 tuned by C_1 and C_2. Each secondary circuit is connected in series with a diode detector and resistor-capacitor load. The loads R_1 and R_2 are in series with each other and the output voltage is the total p.d. developed across these two equal resistors.

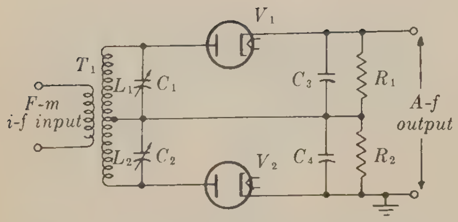


Fig.13-9 Detuned-circuit discriminator.

The i-f signal from the limiter is applied to the primary winding of T_1 which is broadly tuned to the frequency f_0 at the centre of the i-f bandpass. The frequency f_0 is the mean i-f corresponding to the mean r-f of the signal received at the aerial. The secondary circuit L_1C_1 is tuned to a frequency f_1, which is below f_0 by an amount equal to the maximum deviation of 75 Kc/s. L_2C_2 is

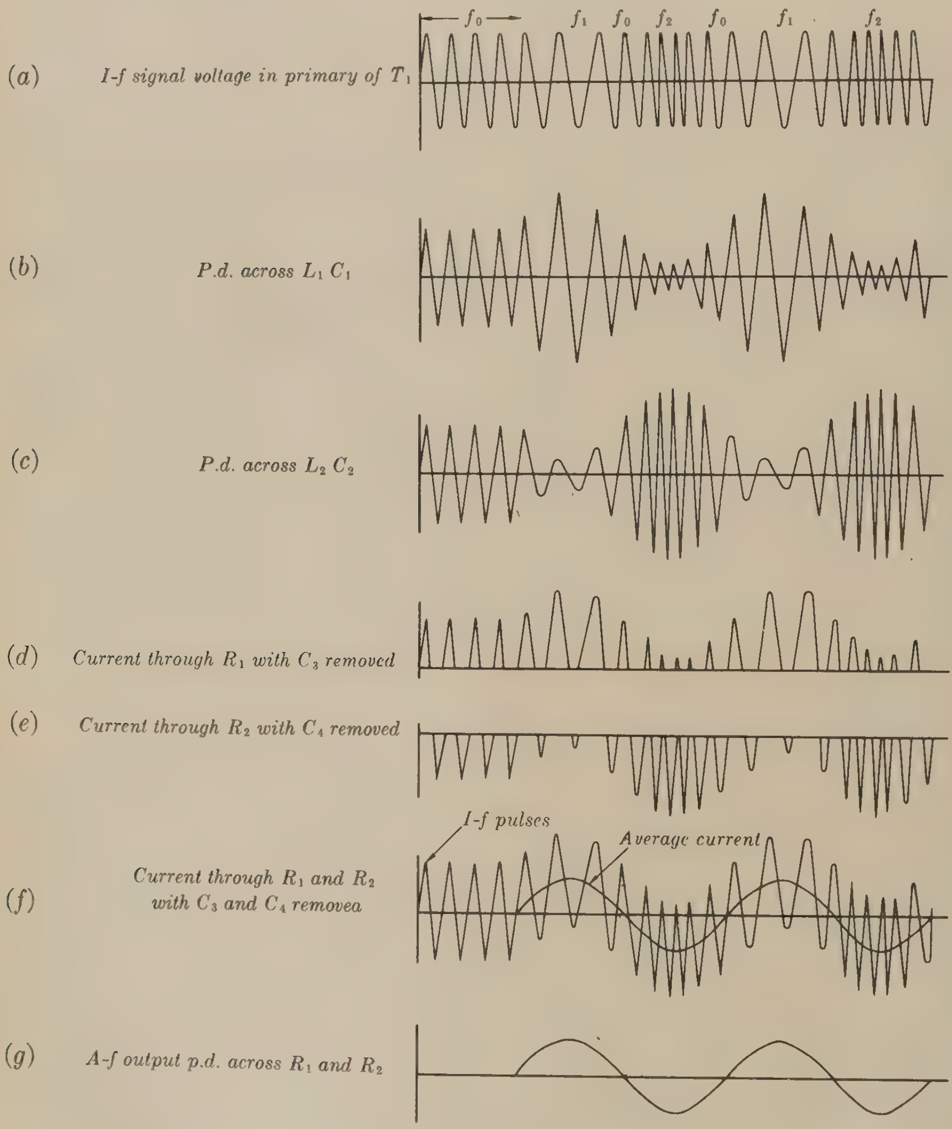


Fig.13-10 Voltage and current waveforms in a detuned-circuit discriminator.

tuned to a frequency f_2 which lies above f_0 by the same amount.

The operation of the circuit is explained with the aid of the waveforms of voltage and current shown in Fig. 13-10. The i-f signal voltage in the primary of T_1 is shown in (a). It is a voltage of constant amplitude. The frequency remains constant at f_0 in the absence of modulation during the period indicated in the graph. When modulation at a single frequency is present the frequency decreases to a minimum at frequency f_1, returns to f_0, increases to a maximum at f_2 and returns to f_0. This completes one cycle of frequency variation. A second complete cycle is also shown.

When the input voltage is at frequency f_0, equal voltages are induced in L_1C_1 and L_2C_2 as shown in graphs (b) and (c). The two diodes V_1 and V_2 are made to conduct alternately as shown in (d) and (e) and the same amount of current flows through each valve. The current through R_1 develops a p.d. across R_1 such that the top end of R_1 is positive to the bottom end. The p.d. across R_2 due to current through V_2 is such that the bottom end is positive to the top end. The capacitors C_3 and C_4 maintain the p.d.'s across these resistors substantially constant from one cycle to the next. Since the p.d.'s are equal and of opposite polarity, the net output voltage is zero as long as the input signal is at the frequency f_0.

When the frequency of the i-f signal falls from f_0 to f_1 the tuned secondary L_1C_1 is operated closer to its resonant frequency. The amplitude of the oscillations in this circuit increases as shown in (b). At the same time L_2C_2 is operated farther from its resonant frequency and the amplitude of its oscillations decreases as shown in (c). V_1 therefore conducts more heavily, graph (d), and V_2 less heavily, graph (e). The p.d. developed across R_1 is therefore greater than the p.d. across R_2 and the resultant output is positive. When the signal frequency rises again from f_1 to f_0 the conditions in the two valve circuits return to those existing at the beginning of the graphs and the output voltage becomes zero again. When the signal frequency rises from f_0 to f_2 the upper circuit L_1C_1 is operated farther from its resonant frequency and the voltage applied to V_1 decreases in amplitude as shown in (b). At the same time, L_2C_2 is operated closer to its resonant frequency and the voltage applied to V_2 increases as shown in (c). As a result, V_1 conducts less heavily and V_2 more heavily than when the input frequency is f_0. The p.d. across R_2 is greater than the p.d. across R_1 and the resulting output is negative as indicated by graph (g). When the frequency of the i-f signal returns again to f_0 the valves conduct equally and the output voltage returns to zero. This completes one cycle of frequency variation of the signal and one cycle of the a-f output.

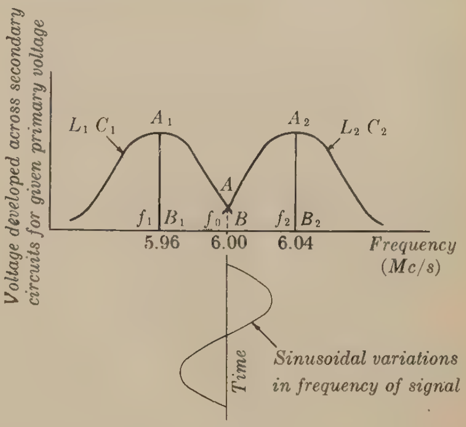


FIG. 13-11 Response curves of secondary tuned circuits of discriminator.

The operation for one particular case is illustrated by the response curves of Fig. 13-11. The curves show the variation in voltage developed across the tuned secondary circuits when an i-f signal of constant amplitude and varying frequency is applied to the primary

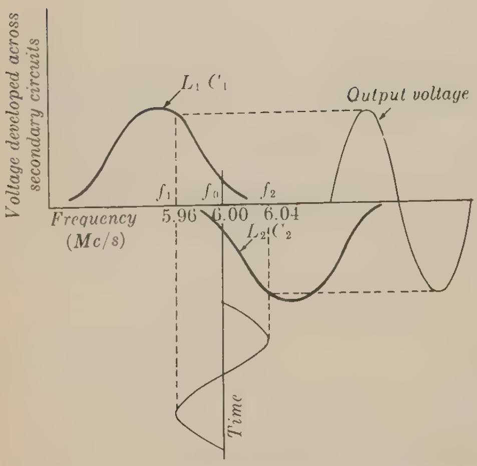


FIG. 13-12 Operating condition of discriminator for no distortion.

circuit of T_1. The input voltage is shown as varying sinusoidally in frequency between the limits 5.96 Mc/s and 6.04 Mc/s about a frequency of 6 Mc/s. When the signal frequency is exactly 6 Mc/s the amplitudes of the voltages in the two secondaries are equal as shown by AB in the diagram. When the signal frequency increases from 6 Mc/s to 6.04 Mc/s the amplitude of the voltage in circuit 1 falls almost to zero while that in circuit 2 rises to A_2B_2. When the signal frequency decreases to 5.96 Mc/s the amplitude of the voltage in circuit 1 rises to A_1B_1 while that in circuit 2 falls to a negligible value.

The operation is also summarized by the diagram of Fig. 13-12, which shows the response curves again for the secondary tuned circuits of Fig. 13-9. In this diagram, however, the response curve of the L_2C_2 tuned circuit is inverted compared to that of the L_1C_1 tuned circuit since the voltage developed across C_3 and R_1 is opposite in polarity to that across C_4 and R_2. As the input frequency varies between 5.96 and 6.04 Mc/s, the response or the output voltage of the discriminator also varies in a corresponding way as shown.

The disadvantage of this type of discriminator is that the two secondary circuits must be exactly balanced if distortion is to be avoided. That is, the coils must be identical and the diodes, resistors and capacitors must be matched. If the curves in Fig. 13-12 were not symmetrical, a sinusoidal variation in frequency of the signal would not produce a sinusoidal output voltage. The result would be the introduction of second harmonic distortion into the output.

In order that the output amplitude be a linear function of the input frequency the following relationships which have been found experimentally should hold between the Q's of the tuned circuits.

$$Q_p = \frac{f_0}{3(f_1-f_0)}$$

and $$Q_s = 2Q_p$$

where Q_p = Q of the primary circuit
Q_s = Q of either secondary circuit
f_0 = carrier frequency in c/s
f_1-f_0 = frequency deviation in c/s.

These relationships show that the primary circuit is much more broadly tuned than the secondary circuits.

C. Coupled-Circuit Type of Discriminator

A more common type of discriminator is shown in Fig. 13-13. It differs in construction from that of Fig. 13-9 in that a single capacitor C_2 is used to tune the secondary circuit to resonance and a capacitor C_3 is used to couple the primary voltage to the centre tap of the secondary circuit. The voltage across L_2 is applied to the diode V_1 in series with its load circuit R_1C_4. The voltage across L_3 is applied to V_2 and its load circuit R_2C_5. The load circuits are connected in series. L_4 is a radio-frequency choke coil to complete the diode circuits and provide a high r-f impedance between the centre tap of the secondary winding and ground. The input to the circuit is the i-f signal from the limiter V_3. Both the primary

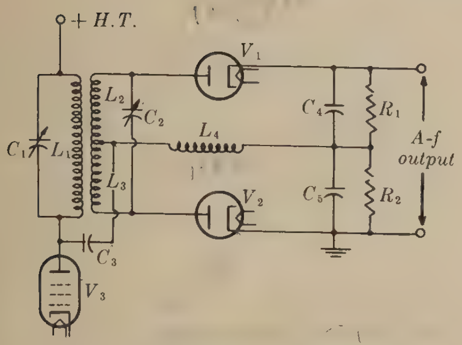


FIG. 13-13 Coupled-circuit discriminator.

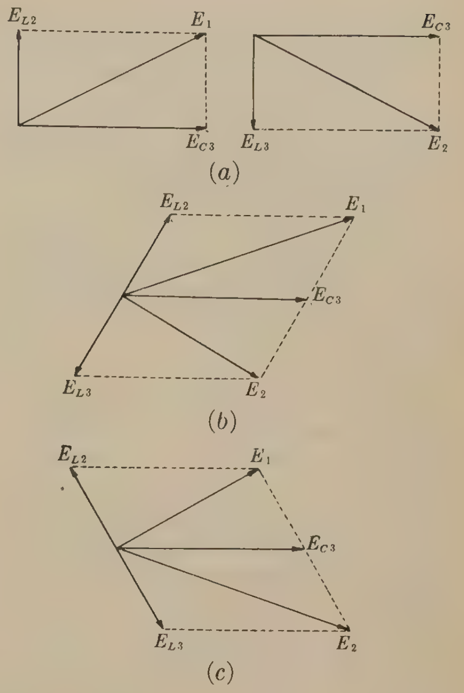


FIG. 13-14 Vector representation of voltages in coupled-circuit discriminator (a) Signal frequency equal to the resonant frequency of the circuit (b) Signal frequency below resonant frequency of the circuit (c) Signal frequency above resonant frequency of the circuit.

and secondary windings of the i-f transformer are tuned to the centre of the i-f band.

The voltage applied to the upper diode is the resultant of two components E_{L2} and E_{C3}. E_{L2} is the voltage induced in the upper half of the secondary winding and E_{C3} is the anode voltage of V_3 applied to the centre tap of the secondary winding by C_3. At resonance E_{L2} leads E_{C3} by 90° as shown in Fig. 13-14 (a). The resultant voltage E_1 is applied to V_1. Similarly, the voltage E_2 which is applied to the lower diode is the resultant of two components E_{L3} and E_{C3}. E_{L3} is 180 degrees out of phase with E_{L2} and therefore lags E_{C3} by 90° as shown. E_1 and E_2 are equal in magnitude when the circuit is operated at the resonant frequency. That is, the anode of V_1 is driven the same amount positive to the cathode during its cycle as the lower anode is made positive during its cycle. Consequently both valves conduct the same amount of current. The fact that the anode voltages are out of phase with each other does not affect the amplitude of the current pulses through each valve. The p.d.'s developed across R_1 and R_2 are steady due to the presence of C_4 and C_5; they are equal and of opposite polarity. Hence the output voltage is zero.

When the frequency of the signal decreases, the voltage E_{L2} leads E_{C3} by less than 90° while E_{L3} lags E_{C3} by more than 90° as shown in Fig. 13-14 (b). The magnitude of E_1 is then greater than the magnitude of E_2. Consequently the anode of V_1 is driven more positive than the anode of V_2 and the p.d. across R_1 is greater than that across R_2. The output voltage is therefore positive.

When the signal frequency increases above the mean frequency, E_{L2} leads E_{C3} by more than 90 degrees and E_{L3} lags E_{C3} by less than 90 degrees, Fig. 13-14 (c). E_2 is therefore greater than E_1 and V_2 conducts more current than V_1. The p.d. across R_2 exceeds the p.d. across R_1 and the output voltage is negative. As the i-f signal undergoes one complete cycle of frequency variation, the discriminator produces one complete cycle of a-f output. The frequency of the

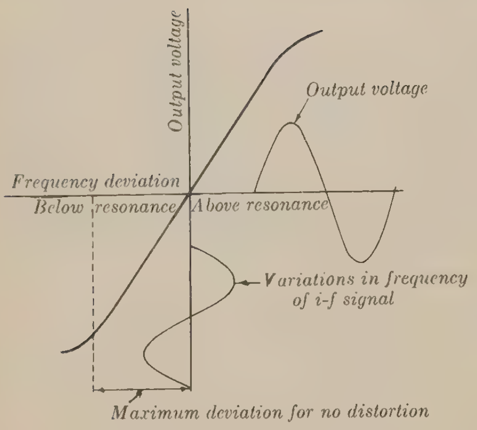


FIG. 13-15 Response curve of f-m discriminator.

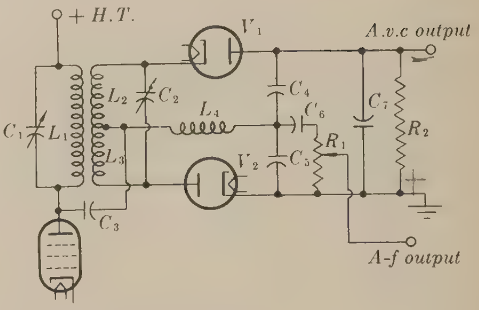


FIG. 13-16 Ratio detector circuit.

output equals the modulating frequency while the amplitude depends upon the amount of frequency deviation in the input signal. The relationship between the frequency deviation of the i-f and the output voltage of such a discriminator is shown in Fig. 13-15. The maximum deviation for no distortion is usually about 75 Kc/s which is the deviation corresponding to 100% modulation in amplitude modulation.

This type of discriminator circuit is superior to the detuned-circuit type of discriminator since the secondary coil need not be accurately centre-tapped. Any slight unbalance can be compensated for by slight detuning of the secondary tuned circuit. The alignment is also easier since there is only one secondary tuned circuit.

D. Ratio Detector

The discriminators described in the preceding paragraphs require that the amplitude of the input signal remain constant to avoid introducing distortion. If the amplitude of the signal increases, both diodes conduct more heavily. As long as the signal frequency is at the centre of the i-f band, the increased output of V_1 is just counteracted by the increased output of V_2. When the signal is above or below the centre frequency, however, one valve conducts more heavily than the other giving rise to a positive or negative output voltage from the discriminator. Under these conditions, the amplitude of the output increases as the amplitude of the i-f signal rises. That is, amplitude modulation of the i-f signal causes amplitude modulation also of the a-f signal. Noise pulses in the input pass through the discriminator and appear in the a-f stages. For this reason, a limiter stage must always precede this type of discriminator.

The *ratio detector* shown in Fig. 13-16 overcomes this difficulty and makes the limiter stage unnecessary. The circuit is similar to that of Fig. 13-13 except that there is a reversal of connections to V_1 and the output circuit is modified as shown. The valves conduct on the same half cycle of input voltage and in the same direction through the load circuit R_2, C_7, C_4 and C_5, causing the upper end of R_2 to be at a negative potential. The time constant of the large electrolytic capacitor C_7 and resistor R_2 is very large, of the order of one quarter second. Hence C_7 charges to a steady voltage which does not vary at even the lowest frequency a-c components. This steady negative voltage across C_7 and R_2 which is proportional to the i-f carrier is used as an a.v.c. voltage. Also, if the amplitude of the i-f voltage applied to this detector circuit rises suddenly for a short

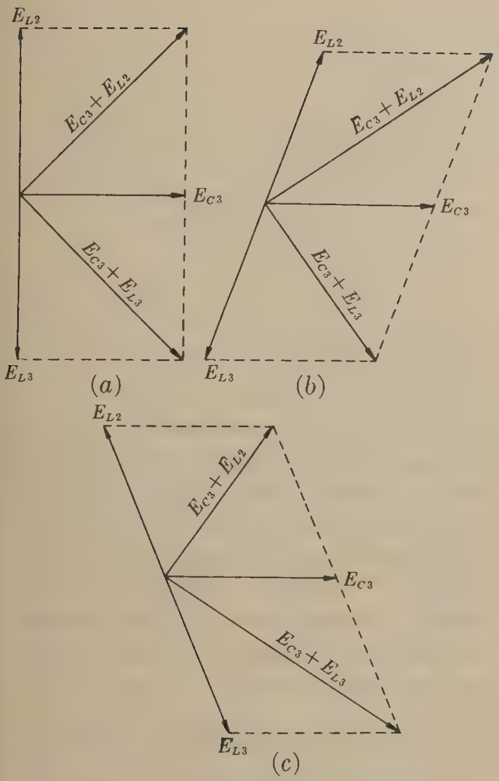


Fig. 13-17 Operation of ratio detector.

period of time due to a burst of static type of interference the voltage across C_7 does not rise because of its large capacitance. Since this steady voltage is applied across C_4 and C_5 in proportion to their capacitances then the voltages across C_4 and C_5 are not changed either by the sudden change in amplitude of the i-f signal. Thus this ratio detector circuit does not respond to amplitude modulation caused by static. It can therefore be used without a preceding limiter circuit.

The voltages across C_4 and C_5 are caused to vary at the a-f rate by the variation in the frequency of the i-f f-m signal. This is explained using the vector diagrams of Fig. 13-17. The voltage applied to the diode V_1 and capacitor C_4 is the vector sum of E_{L2} and E_{C3}. E_{C3} is the a-c component of the output voltage of the preceding stage and is almost entirely developed across the r-f-c L_4. The voltage applied to V_2 and C_5 is the vector sum of E_{L3} and E_{C3}. When the i-f signal is at its resting frequency, E_{L2} and E_{C3} are in quadrature phase as are E_{L3} and E_{C3} as shown in Fig. 13-17 (a). The voltages applied to the two diodes are therefore equal in magnitude. Both diodes conduct every second half cycle when the upper ends of L_2 and L_3 are negative with respect to their lower ends; the current which flows charges C_4 and C_5 with the same polarity and to the same value of voltage. When the i-f signal frequency rises above or falls below the resting frequency, first one or the other of the voltages applied to the diodes increases and decreases from its value at the resting frequency, as indicated in Fig. 13-17 (b) and (c). This occurs because of the change in the relative phases of E_{L2}, E_{L3} and E_{C3} already explained in the previous subsection. Therefore the voltages across C_4 and C_5 change accordingly, the voltage across one increasing when that across the other is decreasing and vice versa. Furthermore since the change in i-f frequency is according to the a-f modulation, the voltages across C_4 and C_5 both vary according to the a-f modulation. However, at all times their sum, the voltage across C_7, must be constant. Hence the a-f output must be taken from across C_4 or C_5 but not from across C_7. In the circuit given, the a-f output is coupled to the succeeding stage by C_6 and R_1 from across C_5. The ratio of the voltages across C_5 and C_4 changes as the i-f frequency changes, giving rise to the name ratio detector.

It is to be noted that while the voltage across C_5 varies because of the change in i-f frequency, it does not vary if the amplitude of the i-f is suddenly changed. That is, the output of this circuit responds to frequency changes of the signal but not to sudden amplitude changes of the signal.

This circuit has two advantages then over the previously described discriminators.

(i) It possesses an inherent ability to reject amplitude modulation of the static type which f-m communication seeks to avoid. No preceding limiter stage is then necessary.

(ii) It develops an a.v.c. voltage.

The ratio detector finds wide application in television as well as in radio communications.

13.6 Methods of obtaining frequency modulation. It is pointed out in section 13.1 that the f-m transmitter differs from the a-m transmitter chiefly in the modulation stages. Two methods of frequency modulating the carrier are in general use. One is known as the *reactance-tube modulating system* and the other as the *phase-shift modulating system.* In the reactance-tube system the r-f carrier is modulated at the oscillator by varying the effective inductance or capacitance of the tuned circuit. In the phase-shift modulating system the r-f carrier is modulated in some stage following the oscillator. This system operates on the principle that a frequency-modulated wave can be obtained from a phase-modulated wave with simple circuit modifications. Methods of effecting frequency modulation are described in the following section and phase modulation is discussed later.

13.7 Reactance-tube system of frequency modulation. The use of a capacitor microphone in the simple transmitter described in section 13.2 is not practicable since it cannot produce sufficient deviation even with the use of frequency multipliers. A circuit which does produce enough frequency deviation and employs the same principle of varying the reactance of an oscillator circuit in accordance with an audio modulation is the reactance-tube modulator. In this case an electronic circuit which behaves like a variable reactance is shunted across the oscillator tuned circuit. The reactance is made to vary by the a-f signal. This variation in reactance causes a similar variation in the frequency of the oscillatory circuit, thereby producing an f-m wave.

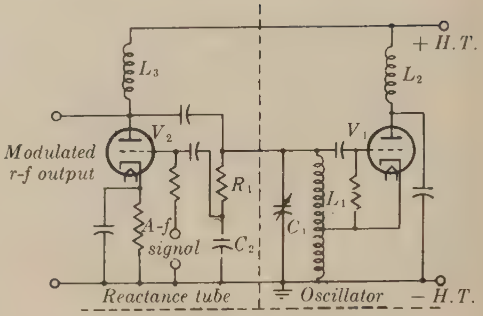


Fig.13-18 Reactance-tube modulator with inductive loading.

A. Inductive Reactance-Tube Circuit

One type of reactance-tube modulator is illustrated in Fig. 13-18. In this circuit, V_1 is a modified Hartley r-f oscillator with L_1 and C_1 the main components of the tuned circuit. Connected across this tuned circuit is R_1 and C_2. The resistance of R_1 is large compared to the reactance of C_2 at the oscillator frequency so that the current in the R_1C_2 branch is very nearly in phase with the tank voltage applied across it. The voltage across C_2 however lags the current by 90°. This voltage which is applied to the grid of V_2, the reactance-tube circuit, causes the anode current of V_2 to vary at the r-f rate. The a-c component of the anode current is blocked from H.T. by the r-f choke L_3 and forced to return to ground through the tank and R_1C_2 combination in parallel. Since the tank circuit impedance is very much less than that of R_1, most of the a-c component of anode current of V_2 will flow in the tank circuit rather than R_1.

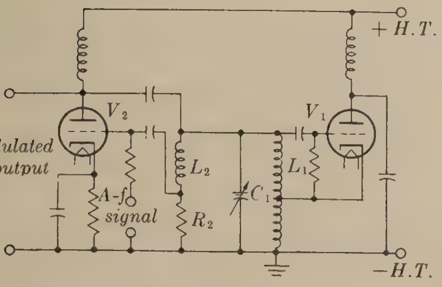


FIG. 13-19 Alternative type of reactance-tube modulator with inductive loading.

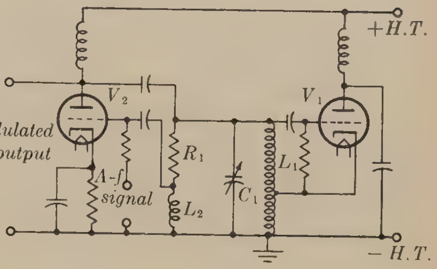


FIG. 13-20 Reactance-tube modulator with capacitive loading.

This current is in phase with the grid voltage and hence the voltage across C_2 and therefore lags the tank circuit current by 90°. Since an injected lagging current in a circuit is equivalent to an inductive reactance in parallel with the circuit, the anode circuit of V_2 appears as an inductance across the tank circuit. The resonant frequency of the tank circuit then depends not only on the values of L_1 and C_1 but also on the value of equivalent inductance across L_1C_1 due to the lagging anode current of V_2 through L_1C_1. The amount of frequency variation of the tank circuit depends on the magnitude of the variation in the anode current of V_2 which can be controlled by the grid bias of V_2. The modulating a-f voltage is applied as bias to V_2 in addition to its cathode bias. This a-f voltage varies the anode current which in turn causes the tank frequency to vary at an audio rate and frequency modulation results.

In the circuit of Fig. 13-19, a similar inductive effect on the tank circuit is obtained by applying to the grid of the modulation valve the voltage across R_2. In this case, the resistance of R_2 is small in comparison with the inductive reactance of L_2. The current in L_2 and R_2 lags the applied voltage from the tank circuit. Hence the voltage across R_2 is a lagging voltage applied to the grid of V_2. The operation of the circuit is otherwise similar to that of Fig. 13-18.

B. Capacitive Reactance-Tube Circuit

A capacitive reactance may be injected into the tank circuit instead of the inductive reactance previously described to produce frequency modulation of the r-f carrier. This requires a modulator anode current that leads the tank voltage. One method of obtaining a voltage at the grid of V_2 which leads the oscillator tank voltage by 90° is shown in Fig. 13-20. In this circuit R_1 and L_2 are connected in series across the tank circuit of the oscillator. The value of R_1 is large in comparison with the reactance of L_2. Thus the current through L_2 and R_1 is practically in phase with the voltage across the tank circuit. Then the voltage across L_2 which leads the current by 90° also leads the tank voltage by 90°. The voltage across L_2 is applied to the grid of V_2. Thus the anode current of V_2 leads the tank voltage by 90° and hence the V_2 circuit acts as a capacitive reactance across L_1C_1, the tuned circuit of the r-f oscillator. As the grid bias varies at an a-f rate the value of the V_2 anode current also varies so that the effective capacitance of the oscillator tank circuit and hence the oscillator frequency varies at the a-f rate.

In the circuit of Fig. 13-21, C_2 and R_1 are in series across the oscillator tank circuit. The reactance of C_2 is large compared to the resistance of R_1. Thus the current through C_2 and R_1 leads the tank

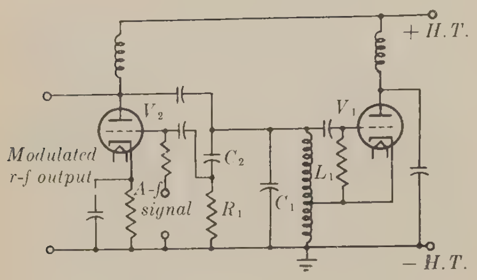


FIG. 13-21 Alternative type of reactance-tube modulator with capacitive loading.

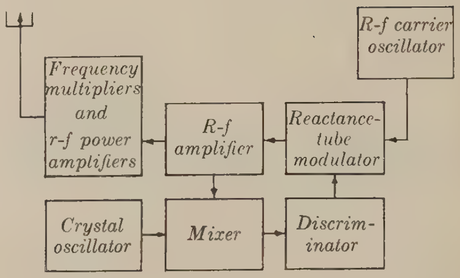


FIG. 13-22 Block diagram of frequency-stabilization circuit.

voltage by almost 90°. The voltage across R_1 also leads the tank voltage by 90°. This voltage is the input to V_2 and causes the anode current of V_2 to lead the tank voltage by 90°. Thus the V_2 circuit acts as a capacitive reactance across the oscillator tank circuit as in the case of Fig. 13-20.

The phase lag or lead of the voltage applied to the modulator valve should be as nearly equal to 90° as possible. If the phase difference is less than 90°, the reactive components injected into the tank circuit are reduced in magnitude; hence the resultant frequency deviation is less.

The reactance-tube modulator is widely used in f-m transmitters. To provide the increased frequency deviation required in the final output such a modulator must be followed by several stages of frequency multipliers to increase the frequency to 8 or 10 times the oscillator frequency.

One disadvantage of the reactance-tube modulator is that the carrier frequency of the oscillator tends to drift. To prevent this the use of frequency-stabilization circuits is required.

13.8 Frequency-stabilization circuit. In an a-m transmitter, the frequency of the oscillator is kept constant by various means such as crystal control, power-supply regulation and temperature compensation. In an f-m transmitter, the frequency of the oscillator is varied by the modulator so that direct crystal control is impossible. The carrier frequency of the oscillator is still subject to change due to mechanical vibration, voltage and temperature variations, etc. The *frequency-stabilization circuit* is designed to counteract drifting of the carrier frequency. A block diagram of the circuit is shown in Fig. 13-22.

The output of the reactance-tube modulator is amplified by the r-f amplifier and passed to the mixer stage. This stage also receives an input of constant frequency from the crystal oscillator. The crystal frequency is made to differ from the carrier frequency by a specified amount. The output of the mixer is a voltage of intermediate frequency equal to the difference between the two mixed frequencies. Usually the i-f is in the neighbourhood of 1500 Kc/s. This signal is passed to the discriminator which is tuned to the i-f. The d-c output voltage of the discriminator is used as bias on the reactance-tube modulator. When the carrier frequency of the oscillator differs from the crystal frequency by 1500 Kc/s, the d-c output of the discriminator is zero and no correcting bias is applied to the modulator. If the carrier frequency drifts from its assigned value, the i-f shifts from 1500 Kc/s. The d-c output of the discriminator is applied as bias to the reactance-tube modulator. The polarity of this voltage is such as to change the reactive current flowing in

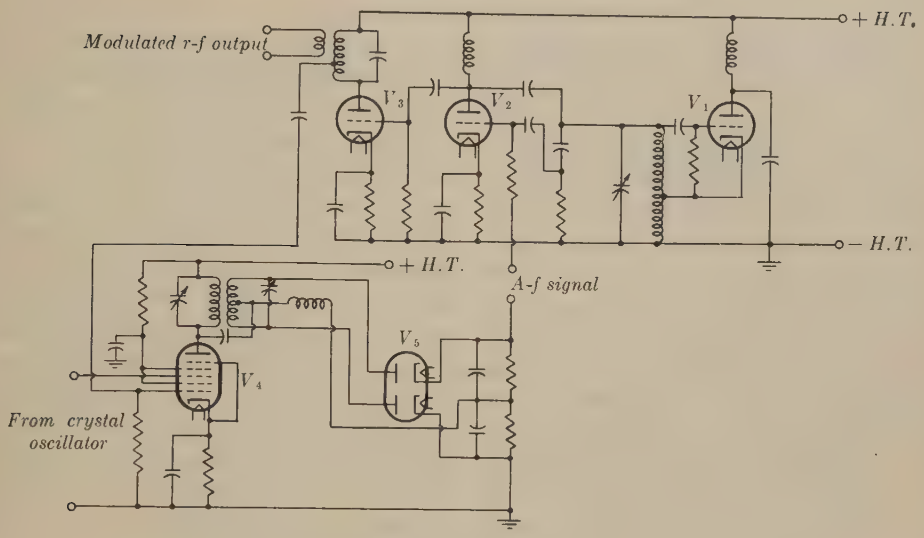


FIG. 13-23 Simple frequency stabilization circuit.

the oscillator tank circuit so that the frequency drift is corrected.

Fig. 13-23 shows a diagram of a simple frequency-stabilization circuit. V_1 and V_2 are the oscillator and reactance-tube modulator stages respectively. Modulation is provided by capacitive loading of the oscillator as in Fig. 13-21. V_3 is the r-f amplifier which provides f-m voltages to the power-amplifier stages and to the mixer valve V_4. This valve also receives the constant-frequency input from the crystal oscillator. The i-f output of V_4 is applied to the discriminator circuit. The d-c output of the discriminator valve V_5 is applied as bias in series with the a-f modulating signal to the grid of the modulator V_2. A particular example will illustrate the operation of the circuit.

Suppose the crystal oscillator is adjusted to operate 1500 Kc/s above the r-f carrier frequency. If the latter frequency tends to increase, the i-f decreases. Since the discriminator is tuned to 1500 Kc/s, the upper diode of V_5 is made to conduct more current than the lower diode. The output of the discriminator is a positive d-c voltage which is applied to the grid of V_2. Consequently the bias on the modulator is reduced and the valve conducts more current. This increases the effective capacitance in parallel with the tank circuit of the oscillator causing its frequency to decrease. Thus, any tendency for the oscillator frequency to drift is counteracted by the frequency-stabilization circuit.

While the above circuit illustrates the principle of operation of frequency stabilization, a more elaborate circuit is usually required for operation at high frequencies. The input to the mixer stage is generally taken from the output of the frequency multipliers or even from the final power amplifier and may therefore be as high as 100 Mc/s. Since it is impossible to operate a crystal at such a high frequency, the crystal oscillator must be followed by several stages of frequency multipliers to bring the reference frequency close to the f-m carrier frequency. The i-f is therefore kept to a comparatively small value and the sensitivity of the discriminator is maintained at a reasonable level.

Frequency stabilization is also accomplished in some f-m transmitters by a frequency-division method and mechanical readjustment of the transmitter frequency. The outputs of the transmitter oscillator and the crystal oscillator are each applied through separate frequency-divider circuits. The arrangement is designed so that when the transmitter carrier frequency is at its correct value the outputs of the two dividing networks are equal in frequency. If the transmitter carrier frequency shifts, the difference in the output frequencies of the dividing networks is arranged to operate a motor which automatically retunes the transmitter master oscillator. The advantage of this more complicated system of frequency stabilization is that the transmitter carrier frequency is returned exactly to its proper value whereas in the first system discussed above, the transmitter carrier frequency is only returned near its proper value.

13.9 Advantages and disadvantages of f-m. There are two main advantages of an f-m system of communication. The first is that the signal-to-noise ratio can be made much greater than in an a-m system. This is due, first, to the fact that most static or man-made interference causes only amplitude variations in the receiver signal; the frequency variations caused by static are small. Receiver noise caused by shot effect and thermal agitation is present in f-m receivers as well as a-m receivers since this type of noise causes frequency modulation as well as amplitude modulation. However the percentage of noise can be reduced by using wide band f-m.

The second advantage of an f-m system is that modulation does not vary the amplitude and therefore the power of the transmitted signal. The transmitter valves, then, may be operated at maximum efficiency at all times, which results in lower operating costs for high-power transmitters, an important factor in commercial broadcasting. Further, since the a-f does not change the amplitude of the carrier appreciably, the a-f power required in an f-m transmitter is much less than in an a-m transmitter.

The greatest disadvantage of f-m is the wide frequency band required for each channel. A wide band f-m must be used in order to obtain any pronounced improvement in signal-to-noise ratio over an a-m system, since the frequency deviations due to noise are then very much smaller than those due to the signal voltage. This necessity of using a wide band restricts f-m to the higher frequency channels where space is available. As a result f-m transmission is limited almost to line-of-sight communication with a maximum dependable range of about fifty miles.

13.10 Sidebands in frequency modulation. The sideband frequencies required for satisfactory transmission of frequency-modulated signals are more difficult to determine than in the case of amplitude-modulated signals. In a-m waves, the sidebands which are introduced by each modulating frequency are $F+f$ and $F-f$ where F is the carrier frequency and f the modulating frequency. In f-m waves, the sideband components depend not only on the frequency but also on the amplitude of the modulation signal. The bandwidth required for satisfactory transmission of f-m signals therefore depends upon the maximum frequency deviation F_d and the frequency of the modulating signal f.

The sideband frequencies in the modulated signal arise from the fact that the sinusoidal carrier wave is distorted during the modulating process. This distortion is illustrated by the waveforms of Fig. 13-24 in which one cycle of a frequency-modulated wave is compared with one cycle of a sinusoidal wave of the same period. In Fig. 13-24 (a) the frequency is gradually decreasing and the negative portion of the modulated wave

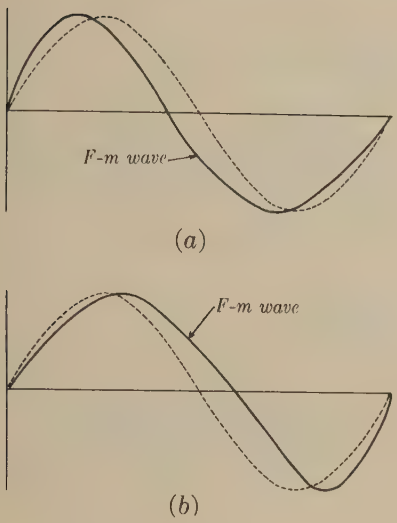


FIG. 13-24 Waveform diagrams to illustrate distortion in an f-m wave (a) Frequency decreasing (b) Frequency increasing.

is of greater duration than the positive portion. In Fig. 13-24 (b) the frequency of the modulated wave is increasing and the positive portion has the greater duration. The distorted wave is equivalent to a sine wave plus sinusoidal components at harmonic frequencies. Mathematical analysis of an f-m wave shows that sideband components of frequencies $F \pm f$, $F \pm 2f$, $F \pm 3f$, etc., are all present. The amplitude of each sideband present depends on the variation in shape of the f-m wave from the sine wave, which in turn depends on the frequency and amplitude of the modulating signal. On a frequency scale, the sidebands are symmetrically placed on either side of the carrier frequency and although there is an infinite number theoretically, in practice the number required for high-fidelity reception is limited.

A. Frequency Spectrum and Bandwidth Requirements

The frequency spectrum of a particular f-m wave is shown in Fig. 13-25. The r-f carrier frequency is 50 Mc/s. The modulating frequency is 5 Kc/s and its amplitude is such as to produce a deviation of 50 Kc/s on either side of the carrier. A mathematical analysis of the waveform shows that fourteen sidebands of appreciable amplitude are produced on either side of the carrier to transmit the signal. Higher order sidebands exist but their amplitudes are so small that they may be neglected. The significant sidebands are plotted to produce the frequency spectrum of Fig. 13-25. The position and height of the lines of the spectrum indicate the frequency and energy level respectively of each of the sidebands. The spectrum shows that most of the energy in the wave is contained in the sidebands and that some of the sidebands contain more energy than the carrier itself. It should be noted that while the frequency deviation is 50 Kc/s on each side of the carrier, the bandwidth required for transmission of the 14 sidebands is 140 Kc/s.

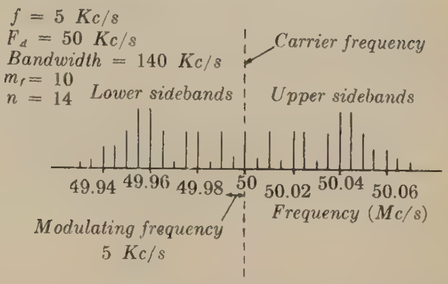


FIG. 13-25 Frequency spectrum of an f-m wave with the characteristics given.

B. Effect of Varying the Amplitude of the Modulating Signal

A change in the amplitude of the modulating signal produces a change in the frequency deviation of the f-m carrier. This results in a frequency spectrum in which the bandwidth and energy distribution are different from those in the original spectrum. If the frequency deviation is reduced from 50 Kc/s to 25 Kc/s, while the frequencies of the modulating signal and carrier are held

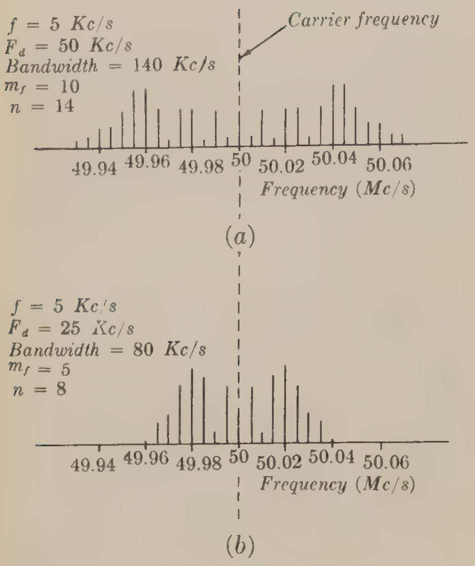


FIG. 13-26 Effect of decreasing frequency deviation with modulation frequency constant (a) Frequency deviation 50 Kc/s (b) Frequency deviation 25 Kc/s.

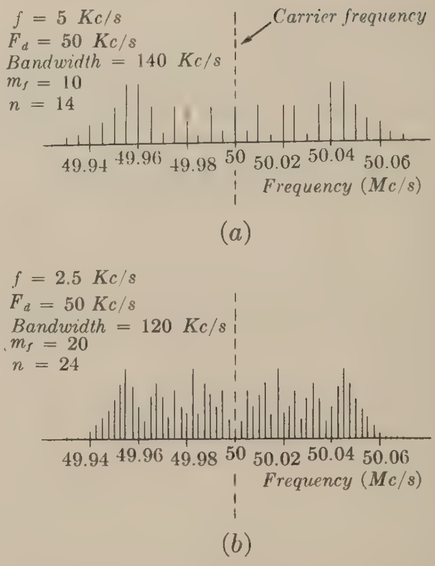


FIG. 13-27 Effect of decreasing modulating frequency with frequency deviation constant (a) Modulating frequency 5 Kc/s (b) Modulating frequency 2.5 Kc/s.

constant, Fig. 13-26 (a) and (b), the number of significant sidebands falls from fourteen to eight. The bandwidth is reduced from 140 Kc/s to 80 Kc/s and the distribution of energy in the sidebands is changed as shown by the spectrum.

C. Effect of Varying the Frequency of the Modulating Signal

If the frequency of the modulating signal is reduced while the deviation is kept constant, the frequency spectrum is altered as shown in Fig. 13-27. A reduction in the frequency of the modulating signal from 5 Kc/s to 2.5 Kc/s results in an increase in the number of significant sidebands from fourteen to twenty-four. The bandwidth is reduced however from 140 Kc/s to 120 Kc/s.

C. Deviation Ratio

The frequency spectrum is dependent upon the frequency deviation F_d and the frequency of the modulating signal f as shown in the preceding paragraphs. Of equal significance is the ratio of the maximum deviation to the maximum modulation frequency. This ratio equals $\frac{F_{d\,max}}{f_{max}}$ and is called the *deviation ratio* or *frequency-modulation index* represented by the symbol m_f. It determines the number of significant sidebands in the spectrum and the distribution of energy in the sidebands. Two waves with different deviations and modulating frequencies but with the same deviation ratio of 5 are shown in Fig. 13-28. The number of significant sidebands and the distribution of energy in the sidebands is the same in each case. In general, the greater the deviation ratio, the greater the number of significant sidebands present in the transmitted signal as shown by the table at the top of page 13:19.

If the deviation ratio is small, a large proportion of the energy is contained in the carrier. As the deviation ratio increases the proportion of energy in the sidebands increases and the energy in the carrier periodically falls to zero.

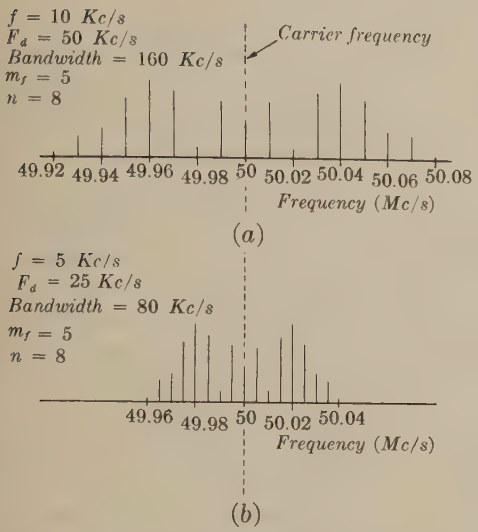


FIG. 13-28 Frequency spectra with constant deviation ratio of five (a) Frequency deviation 50 Kc/s, modulating frequency 10 Kc/s (b) Frequency deviation 25 Kc/s, modulating frequency 5 Kc/s.

13.11 Comparison of frequency modulation and amplitude modulation.

A comparison of frequency modulation and amplitude modulation is given in the table on page 13:20.

13.12 Phase modulation. Frequency modulation with a reactance-tube modulator is accomplished by varying the frequency of the oscillator directly. This system requires extensive frequency stabilization. The system of phase-shift modulation eliminates the necessity of stabilization circuits by accomplishing frequency modulation indirectly and making use of a crystal oscillator. A constant-frequency carrier is phase modulated by the a-f signal and converted to a frequency-modulated wave by a succession of frequency multipliers. The following example illustrates the method. Three cycles of a wave of constant frequency and amplitude are shown in the interval AB of Fig. 13-29 (a). During the following equal interval BC, the phase of the wave is gradually retarded by 45° so that $3\frac{1}{8}$ cycles are produced. The frequency of the wave train is then multiplied to 40 times the original frequency as represented in Fig. 13-29 (b). The interval AB therefore contains 120 cycles while the interval BC contains 125 cycles. The phase modulation introduced in (a) has resulted in frequency modulation in (b). In practice, the total phase deviation in the modulator stage must be kept to about 30° in order to produce linear modulation. A frequency multiplication of several thousand times must be introduced to produce a frequency deviation comparable with f-m. To prevent the carrier frequency from becom-

DEVIATION RATIO $\frac{F_{dmax}}{f_{max}}$	NUMBER OF SIDEBANDS
1	3
2	4
3	5
4	6
5	8
6	9
7	10
8	11
9	12
10	14

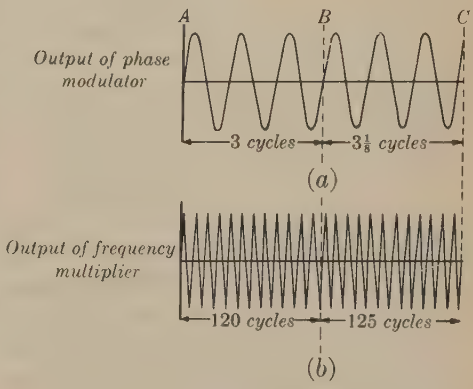


FIG. 13-29 Phase-modulation method of frequency modulation (a) Wave modulated by phase shift of 45° (b) Frequency increased 40 times to show frequency modulation.

Comparison of Amplitude and Frequency Modulation

CHARACTERISTIC	AMPLITUDE MODULATION	FREQUENCY MODULATION
Amplitude of carrier wave	Varies with the degree of modulation	Remains constant during modulation
Amplitude of modulating voltage	Determines the variation in amplitude of the transmitted wave; the stronger the audio signal, the greater the variation in amplitude of the transmitted wave	Determines the deviation in frequency from the carrier frequency; the stronger the audio signal, the greater the frequency deviation
Frequency of modulating voltage	Determines the rate of change of the amplitude of the carrier wave	Determines the rate at which the carrier frequency changes between its upper and lower limits
Width of effective band transmitted	Determined by the frequency of the modulating signal	Determined by the amplitude and frequency of the modulating signal
Modulating power	Power required for anode modulation depends on the required degree of modulation; it is one half of the anode power input to the modulated stage for 100% modulation	Power required for f-m is negligible; amount required is only enough to supply the anode loss in the modulator valve
Carrier power	The final amplifier must be capable of supplying four times the carrier power on 100% modulation peaks	The final amplifier need only be capable of supplying the carrier power
Frequency limitation	Amplitude modulation is practicable on almost any radio frequency	Frequency modulation is limited to the higher frequencies of the order of 20 Mc/s and above

ing excessive, the modulated wave is heterodyned with an oscillator output of constant frequency and the resulting beat frequency obtained. In this way the frequencies of the carrier and sidebands are reduced while the frequency deviation is maintained. The process of increasing the frequency by multipliers and reducing it by heterodyning is repeated as often as required to obtain the desired deviation and final carrier frequency. While this system requires a great many individual circuits, the stages may be operated at sufficiently low power

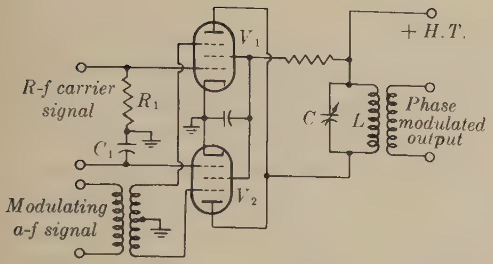


FIG. 13-30 Phase modulator circuit.

to justify the method, especially in view of the high frequency stability obtainable with a crystal oscillator.

A. Phase Modulator

A simple phase modulator circuit is shown in Fig. 13-30. V_1 and V_2 are two Class C r-f amplifiers connected so that their output currents are fed in parallel to an LC circuit tuned to the carrier frequency. The control grids are driven by the r-f carrier voltage from the crystal oscillator. This voltage is applied across the phase-shifting circuit R_1C_1. As a result, the grid voltage applied to V_2 lags that applied to V_1 by 90° and so the anode currents i_1 and i_2 are 90° out of phase with each other. The resultant r-f current i to the LC circuit is the vector sum of the two components i_1 and i_2. The a-f modulating signal is applied in push-pull to the suppressor grids of V_1 and V_2 so that at any instant equal and opposite voltages act on these grids as shown in Fig. 13-31 (a) and (b). At the beginning of the a-f cycle t_0, the suppressor voltages are equal and so the anode currents i_1 and i_2 are equal. The resultant current i leads i_2 by 45° as shown in the first vector diagram of Fig. 13-31 (c). At time t_1 the voltage on the suppressor grid of V_1 is a positive maximum and that on the suppressor grid of V_2 a negative maximum. Hence i_1 is a maximum and i_2 a minimum. Thus i leads i_2 by more than 45° as shown in the second vector diagram. At time t_2 conditions are the same as at t_0 and the phase difference between i and i_2 is 45° again. At time t_3 the voltage on the suppressor grid of V_1 is a negative maximum and that on the suppressor grid of V_2 is a positive maximum. In this case i_2 is a maximum and i_1 a minimum. Hence i leads i_2 by less than 45° as shown in Fig. 13-29 (c). At time t_4 the conditions are the same as at t_0 and the phase difference between i and i_2 is 45° again. Therefore during one cycle of the a-f signal, the phase of i relative to i_2 undergoes one cycle of changes and so the phase of the output tank voltage undergoes one cycle of changes. The greater the amplitude of the modulating voltage, the greater is the phase deviation.

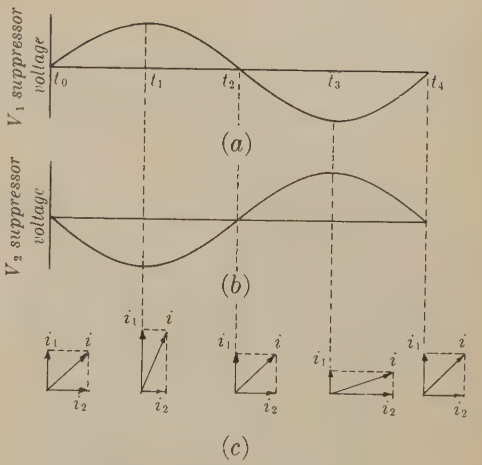


FIG. 13-31 Operation of phase modulator (a) A-f voltage applied to suppressor grid of V_1 (b) A-f voltage applied to suppressor grid of V_2 (c) Vector diagrams of current to tank circuit.

The r-f output voltage has almost constant amplitude since, when one valve conducts more current, the other conducts less current. The small variations in amplitude of the output voltage may be eliminated by the use of a saturated amplifier following the modulator.

B. Sidebands in Phase-Modulated Waves

In both f-m and p-m, the sideband

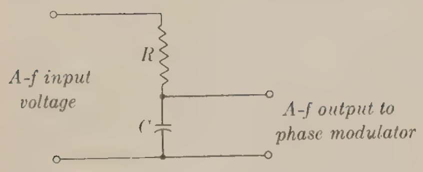


FIG.13-32 Amplitude-correcting network for p-m circuit.

frequencies present are theoretically infinite in number and have values $F \pm f$, $F \pm 2f$, $F \pm 3f$, etc., where F is the frequency of the carrier and f is the modulating frequency. In f-m the amplitudes of the sidebands are inversely proportional to the modulating frequency for a given amplitude of modulating voltage. In p-m, however, the amplitudes of the sidebands for a given amplitude of modulating voltage are independent of frequency. If it is desired to convert the p-m wave to an f-m wave, the amplitude of the sidebands produced in the p-m circuit must be made to decrease as the frequency of the modulating voltage increases. A simple correcting network which accomplishes this is shown in Fig. 13-32. The a-f modulating voltage is applied across R and C in series. The a-f voltage developed across C is applied as the input to the phase modulator. For low frequencies, the reactance of C is high and a large proportion of the input voltage is developed across C. For high frequencies, the reactance of C is reduced and a smaller proportion of the input voltage is developed across C. By making R large in comparison with the reactance of C, the amplitude of the a-f input to the modulator is made inversely proportional to the modulating frequency. In this way the output of the phase modulator may be made to resemble a true f-m voltage.

C. Detection of Phase-Modulated Waves

In the preceding paragraphs the method of producing an f-m wave from a phase-modulated wave is discussed. If this system is employed, the operation of the receiver is the same as that for detecting f-m waves produced by the reactance-tube modulation method. It is possible, however, to transmit phase-modulated waves without conversion to frequency modulation characteristics. In this case, a wider bandwidth is required to accommodate the high frequencies in the modulation signal and a correcting network is required in the receiver. This network is similar to that used in the transmitter to produce f-m waves by the phase-shift modulation process. If the phase deviation of the p-m wave is small, the conversion to an a-m wave which can be detected may be accomplished in a manner similar to the phase-modulation process described previously. The p-m wave is combined with a carrier produced in the receiver. This auxiliary carrier is synchronized 90° out of phase with the carrier component of the p-m wave before being combined with it. The result is an amplitude-modulated wave which can be detected by the ordinary a-m detector.

13.13 Pulse modulation. An alternative method to the transmission of intelligence by a continuous wave, is the transmission by a series of short carrier-wave pulses. Five types of pulse modulation are described in the following paragraphs. These include *pulsed amplitude modulation* (*p-a-m*), *pulsed frequency modulation* (*p-f-m*), *pulse width modulation* (*p-w-m*), *pulse time modulation* (*p-t-m*) and *pulse code modulation* (*p-c-m*).

A. Pulsed Amplitude Modulation

Fig. 13-33 shows the waveforms of the a-f modulating voltage and r-f output voltage of a pulsed amplitude transmitter. The waveforms are similar to those produced in continuous wave modulation except that the transmitter is periodically turned on and off. The

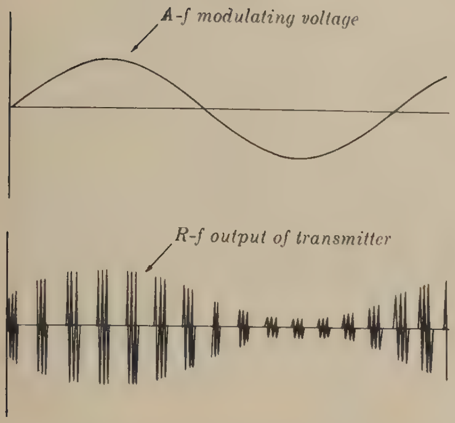


FIG. 13-33 Pulsed amplitude modulation.

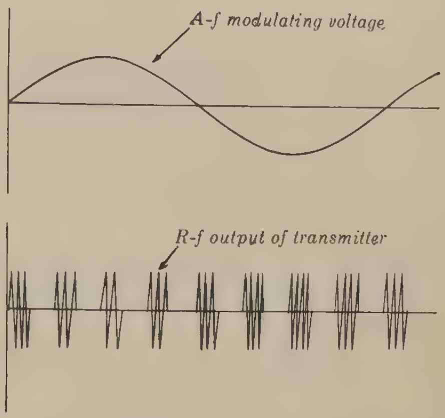


FIG. 13-34 Pulsed frequency modulation.

pulses are equally spaced and of equal duration but vary in amplitude. Usually, the r-f carrier is in the microwave band and the duration of the pulses of the order of one to three microseconds. The pulse repetition rate should be at least twice the highest modulating frequency. An example will serve to illustrate two advantages of this type of modulation. A transmitter produces r-f pulses of two microseconds duration at a repetition rate of 10 Kc/s. One pulse is then produced every 100 microseconds. The transmitter is in operation for $\frac{1}{50}$ of the total transmitting time. The average power of the transmitter is only $\frac{1}{50}$ of the peak power and the transmitting circuits may be driven far above their normal power ratings for continuous operation.

A second advantage of transmission by pulses is the increased signal-to-noise ratio of the system. This also arises from the fact that the transmitter is in operation for only a small fraction of the total transmitting time. The receiver may therefore be silenced for the comparatively long interval between pulses. Noise generated in the receiver and entering by way of the aerial is ineffective during the intervals when the receiver is silenced. As there is no loss in signal and a reduction in noise output, the signal-to-noise ratio is increased.

The detector circuits in the receivers of pulsed systems are somewhat different from those used in continuous-wave systems. The p-a-m detector maintains the output voltage at the peak value of the pulse until the next pulse arrives. This permits the output to fluctuate at the a-f rate but not at the pulse repetition rate.

B. *Pulsed Frequency Modulation*

An f-m wave may also be transmitted as a series of pulses instead of a continuous wave as shown in Fig. 13-34. In this case, the amplitude of the pulses remains constant and the frequency of the carrier is made to vary from pulse to pulse by the modulator. As with p-a-m, the peak power of the transmitter is much greater than the average power and a higher signal-to-noise ratio is attainable if the receiver is silenced during the interval between pulses.

C. *Diplexing and Multiplexing*

It is possible to transmit a series of r-f pulses which are amplitude modulated and frequency modulated independently as shown in Fig. 13-35. Two separate sources of intelligence called *channels* pro-

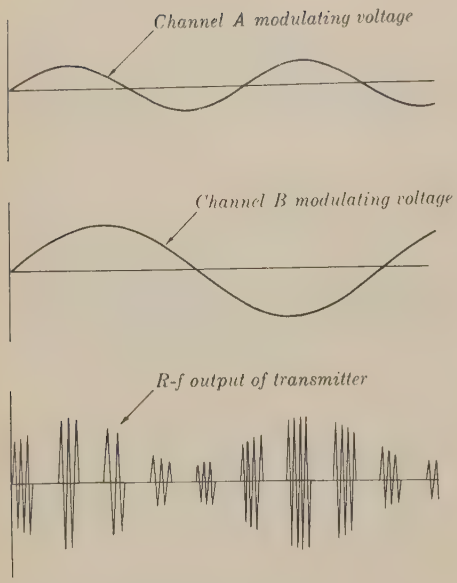

Channel A modulating voltage
Channel B modulating voltage
R-f output of transmitter

FIG.13-35 Diplexing with pulsed a-m and f-m.

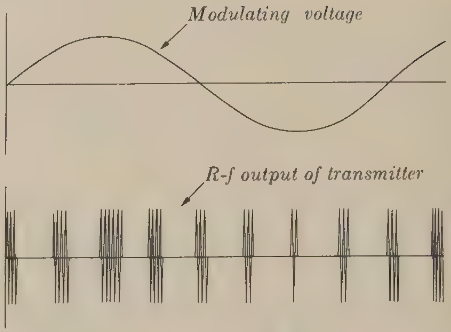


FIG.13-36 Pulse width modulation.

duce a-f signals which are independent of each other in frequency and amplitude. The signal from channel *A* is used to modulate the amplitude of the r-f pulses while the signal from channel *B* is made to modulate the frequency. The intelligence from each type of modulation is recovered and separated in the receiver by two circuits one of which responds to the a-m only and the other to the f-m only. This system of transmission is called *diplexing* and since only one transmitter is used, there is a saving in power, time and equipment.

The advantages afforded by diplexing are further extended by utilizing the interval between r-f pulses to transmit additional signals. Communication by this method is referred to by the terms *multichannel communication* or *multiplexing*. The types of modulation described in the following paragraphs are well suited to multiplexing.

D. Pulse Width Modulation

In pulse width modulation, the transmitter produces r-f pulses of constant

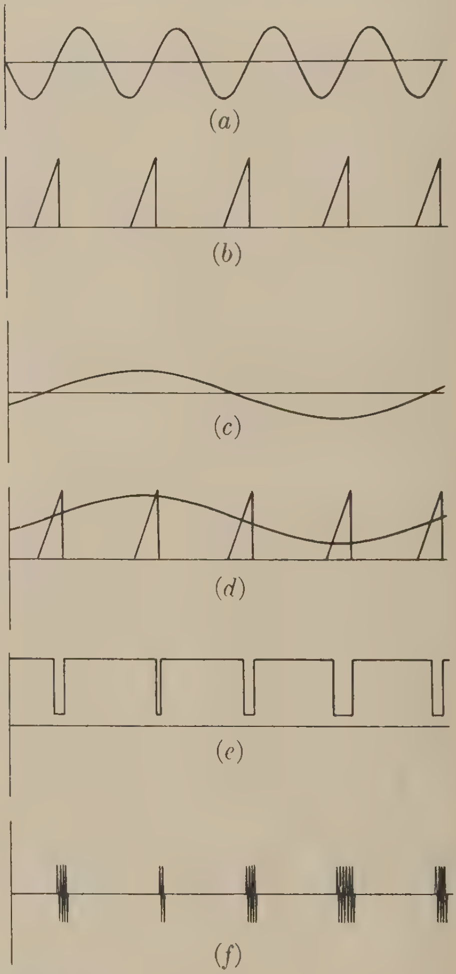


FIG.13-37 Voltage waveforms of a pulse-width-modulation transmitter (a) 10 Kc/s sinusoidal voltage (b) 10 Kc/s sawtooth voltage (c) A-f modulating voltage (d) Combination of sawtooth and modulating voltages (e) Modulation envelope (f) R-f output of transmitter.

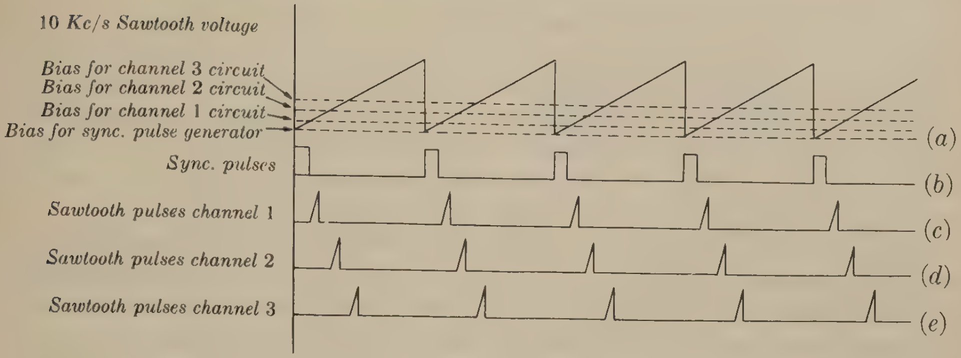

10 Kc/s Sawtooth voltage
Bias for channel 3 circuit
Bias for channel 2 circuit
Bias for channel 1 circuit
Bias for sync. pulse generator
Sync. pulses
Sawtooth pulses channel 1
Sawtooth pulses channel 2
Sawtooth pulses channel 3
(a)
(b)
(c)
(d)
(e)

FIG. 13-38 Production of sync. pulses and sawtooth pulses for multiplexing with pulse width modulation.

amplitude and frequency but of varying duration. The duration or width of the pulses is varied in accordance with the amplitude of the modulating signal while the frequency of the pulse variations is made equal to the modulating frequency (Fig. 13-36).

The operation of the modulator and transmitter is illustrated by the waveforms of Fig. 13-37. The pulse repetition frequency is fixed at about 10 Kc/s by an oscillator which produces a sinusoidal output as shown in (a). This output is used to trigger a sawtooth generator which produces a positive-going sawtooth waveform as shown in (b). The a-f modulating voltage shown in (c) is applied to the cathode of the modulator valve and the sawtooth voltage to the grid of the same valve. Current flows through the modulator when the grid is made sufficiently positive or the cathode sufficiently negative. This condition is indicated in (d) as the time when the sawtooth waveform is above the sinusoidal waveform. When the valve conducts, the anode potential falls producing negative pulses of varying width. These pulses are amplified and squared to form the modulation envelope shown in (e). The modulator then turns the transmitter on for the duration of each pulse producing the r-f output shown in (f).

Pulse width modulation is readily adaptable to multi-channel operation as illustrated in Fig. 13-38. The system handles eight channels and operates in the 6-7 cm. band. Synchronizing pulses are transmitted with the carrier to enable the receiver to separate the channels. The sync. pulses have a duration of 20 microseconds and are transmitted at a constant repetition rate of 10 Kc/s or one pulse every 100 microseconds. Thus the interval between sync. pulses is 80 microseconds. As the average duration of the modulated pulses is 3.5 microseconds, there is sufficient time in the interval to transmit 8 pulses corresponding to 8 separate channels. The intelligence of channel 1 is carried by the first pulse in each interval following the sync. pulse. Channel 2 is made up of the second pulse in each interval and so on.

The process of multiplexing the channels is carried out in the transmitter as follows. The output voltage of the sine wave oscillator described above is used to trigger a sawtooth generator. A positive-going voltage which has the same frequency as the oscillator and which has no steady period is produced as shown in Fig. 13-38 (a). This voltage is made to trigger a sync. pulse generator which produces a square wave of voltage at the beginning of each cycle of the sawtooth waveform as shown in Fig. 13-38 (b). The positive-going voltage of

Fig. 13-38 (a) is also applied to a series of biased-off sawtooth generators, one for each channel. The bias on each of these circuits is somewhat greater than the bias on the preceding one in the series. Each sawtooth generator is triggered when the positive-going voltage overcomes the bias. As a result, the circuits operate successively and produce sawtooth voltage pulses as shown in Fig. 13-38 (c), (d) and (e). While only three channels are indicated in the diagram, it is evident that several more may be accommodated without causing interference among the channels. Each series of pulses may be modulated by a different a-f in the manner described for Fig. 13-37.

Finally, the outputs of the individual modulators are combined to form a series of square pulses. This waveform when amplified is used to modulate the r-f wave train of the transmitter.

At the receiver, the signal is sent to each of the channel units as well as to an integrator which separates the sync. pulse from the channel pulses. The integrator is a series RC circuit with its time constant adjusted to make it respond to the sync. pulses but not to the shorter channel pulses. The integrator triggers a sawtooth generator the output of which operates a gate valve in each of the channel units. These gate valves have a successively greater bias from one channel unit to the next so that each gate valve produces a positive gate pulse of about 8 microseconds duration in succession during the interval between sync. pulses. Although the signal is passed to one grid of the first amplifier in each unit, none of these amplifiers conducts until they receive their gate pulse. Thus unit one amplifies only the first modulated pulse following each sync. pulse. None of the other signal pulses affects this amplifier since its gate pulse occurs only during the reception of the first signal pulse. Similarly unit two is gated on during the reception of the second signal pulse following each sync. pulse and is inoperative while the other channel pulses are received. In this way, the pulses for each unit are separated, each operating its respective unit.

In each channel unit, the signal pulses pass through suitable low-pass filters, with a cut-off at about 4 Kc/s, and the modulation frequencies are recovered and amplified before being passed to appropriate output circuits.

Such a system finds best application in telephone systems, where multiplexing is a necessity in view of economy of time and equipment and where the fidelity obtainable with only 4 Kc/s. a-f band-pass is sufficient for reproduction of voice frequencies.

E. Pulse Time Modulation

In each of the pulse modulation systems so far considered, some characteristic of the pulse itself has been altered in accordance with the audio modulating signal. A transmission by pulse time modulation consists of a series of r-f pulses of constant amplitude, duration and carrier frequency but with variable timing. The time interval between successive pulses is made to vary in accordance with the instantaneous amplitude of the modulating signal. The frequency of the pulse variations is governed by the frequency of the modulating voltage.

Pulse time modulation was developed mainly for multi-channel operation. It has been used in radio and telephone communication of the coaxial cable type, u-h-f broadcasting and television sound channels.

A number of advantages of p-t-m over other types of transmission are as follows.

(i) The signals are of the simplest type. They are short pulses of constant shape and duration with variable timing.

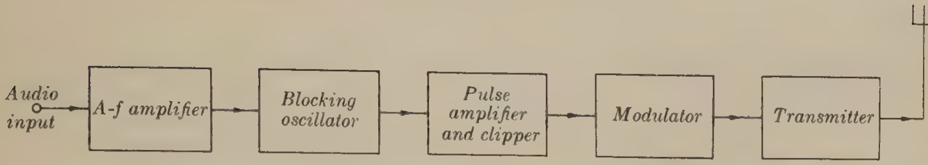


FIG. 13-39 Block diagram of a single-channel pulse-time-modulation transmitter.

(ii) With this type of modulation, parasitics can be considerably reduced.

(iii) The signal-to-noise ratio of a p-t-m system is greater than that of an a-m or f-m system under comparable conditions.

(iv) Noise can only affect the receiver for the short time during which the pulse is being received; the receiver may be silenced during most of the period between pulses.

(v) The pulses may be given additional characteristics that enable them to be separated from noise.

(vi) The cumulative distortions of amplifiers in a-m and f-m systems are not encountered in p-t-m.

(vii) The total bandwidth requirement of a p-t-m system is independent of the number of channels used. This is in contrast to other types of multi-channel operation based on frequency selection in which the bandwidth required for satisfactory transmission increases with the number of channels employed.

A p-t-m transmitter designed for single-channel operation is illustrated by the block diagram of Fig. 13-39. The a-f signal is amplified by an audio amplifier which has a flat response over the operating range from approximately 300 c/s to 2 Kc/s. Its output is fed to a synchronized blocking oscillator of the type described in Chapter XV. The blocking oscillator produces a series of single-swing oscillations of about one microsecond duration in each of its positive and negative half cycles. In the absence of any modulating signal from the a-f amplifier, these pulses are generated at a constant frequency of about 10 Kc/s as indicated in Fig. 13-40 (a). The modulating signal is applied to the blocking oscillator to modulate the frequency of the pulses as shown in Fig. 13-40 (b) and (c). The modulated pulses are amplified and the negative portions clipped off by the following stage. The result is a series of positive square pulses of constant amplitude and duration but of varying time interval as indicated by Fig. 13-40 (d). These pulses are further amplified by the modulator which turns the transmitter on for the duration of the pulse. The transmitter operates at a

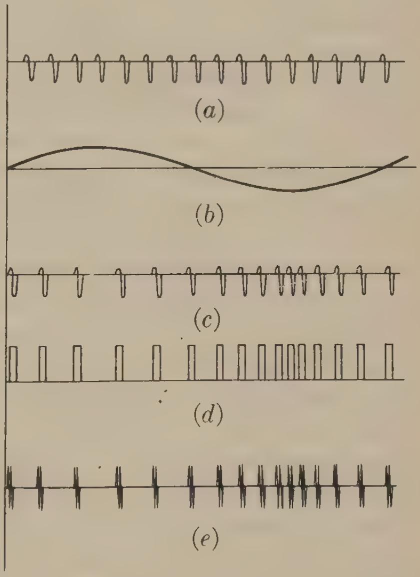


FIG. 13-40 Voltage waveforms produced by p-t-m transmitter circuits (a) Unmodulated output voltage of blocking oscillator (b) A-f modulating signal (c) Modulated output voltage of blocking oscillator (d) Output voltage of modulator (e) R-f output of transmitter.

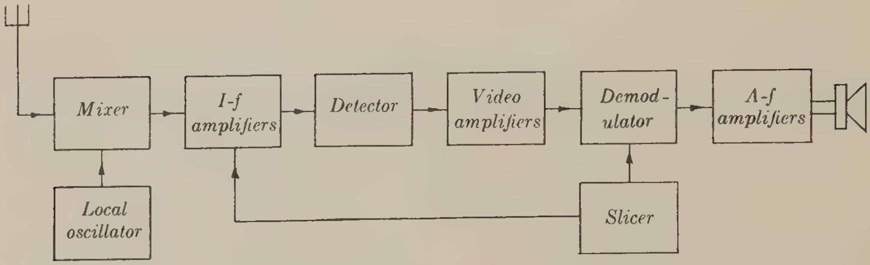


FIG.13-41 Block diagram of a single-channel p-t-m receiver.

frequency of 3000 Mc/s, producing r-f pulses of one microsecond duration and at an average repetition rate of 10 Kc/s, Fig. 13-40 (e).

A superheterodyne receiver suitable for the reception of the single channel p-t-m signals of the above transmitter is illustrated by the block diagram of Fig. 13-41. The aerial signal at 3000 Mc/s is mixed with the output of the local oscillator at 3060 Mc/s to provide an i-f of 60 Mc/s. The i-f signal is amplified by a series of i-f amplifiers and applied to the second detector. The output of the detector is the video signal which consists of a series of square pulses similar to those used to modulate the transmitter. After further amplifications, the video signal is fed to the demodulator which reproduces the a-f signal. The waveforms of the various signals developed throughout the receiver are shown in Fig. 13-42.

The video signal which is passed to the demodulator is also fed to a *slicer* circuit which removes a slice from the pulse at the time of steepest rise and fall. This voltage is used to gate the receiver to remove all noise between pulses.

The demodulator is shown in simplified form in Fig. 13-43. It is essentially a sawtooth generator in which the time constant C_2R_2 is made to correspond to that of the blocking oscillator in the transmitter. The charging capacitor C_2 charges exponentially through R_2 and is discharged through the valve when a positive pulse is received at the grid. The input to the circuit is the video output of the detector. The variation in the period between pulses produces a corresponding variation in the amplitude of the sawtooth voltage across C_2. This output is passed through a low-pass filter which retains the a-f component of voltage and filters out the higher frequency components which are not required. Voltage waveforms which illustrate the operation of the demodulator are shown in Fig. 13-44.

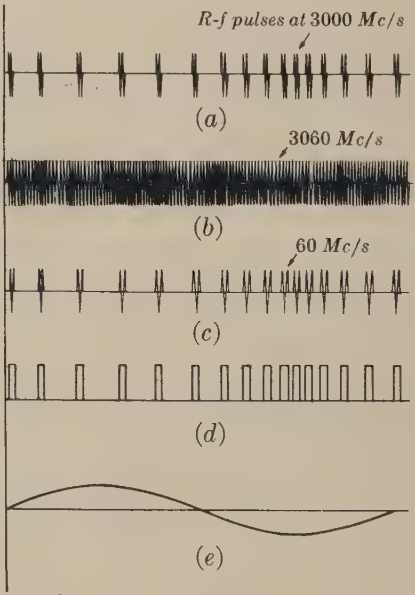


FIG.13-42 Voltage waveforms produced in p-t-m receiver (a) R-f input signal from aerial (b) Local-oscillator output voltage (c) I-f signal applied to detector (d) Video output of detector (e) A-f output of demodulator.

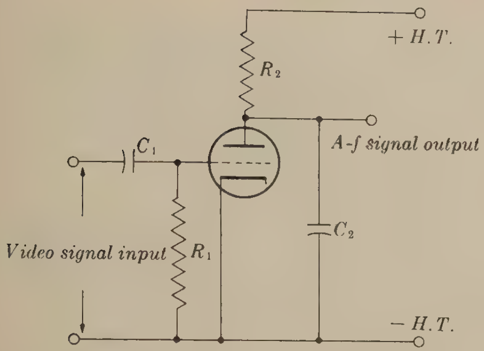


Fig. 13-43 Simplified form of p-t-m demodulator.

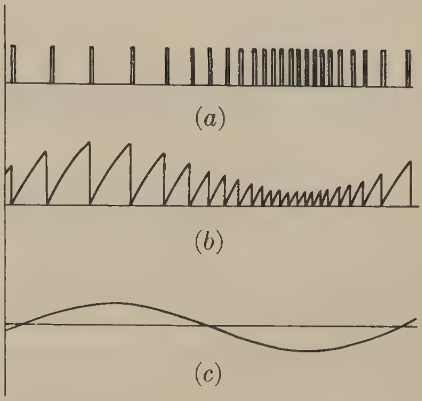


Fig. 13-44 Voltage waveforms produced by p-t-m demodulator (a) Video signal input to demodulator (b) Voltage produced across C_2 (c) Filtered a-f output voltage.

The simplified demodulator of Fig. 13-43 introduces too much distortion to provide satisfactory operation in the p-t-m receiver. A more elaborate demodulator circuit is shown in Fig. 13-45. The sawtooth generator circuit V_1 is followed by a cathode follower V_2. The sawtooth voltage across C_2 is also developed across the cathode load components R_3C_3. This signal is applied through V_3, a bi-directional, pulse-operated triode switch, to C_6 a small *viewing* capacitor. The positive pulses which make up the video input to V_1 are also applied through T_1 as positive pulses to both grids of V_3. These pulses develop automatic bias across R_4 and R_5 which keeps the valve non-conducting except for the duration of each pulse. At the instant when the sweep capacitor C_2 attains its peak voltage, the positive pulse applied to the grids of V_3 permits either triode to conduct and allows C_6 to charge or discharge to the same voltage as C_3. If C_3 is charged to a higher voltage

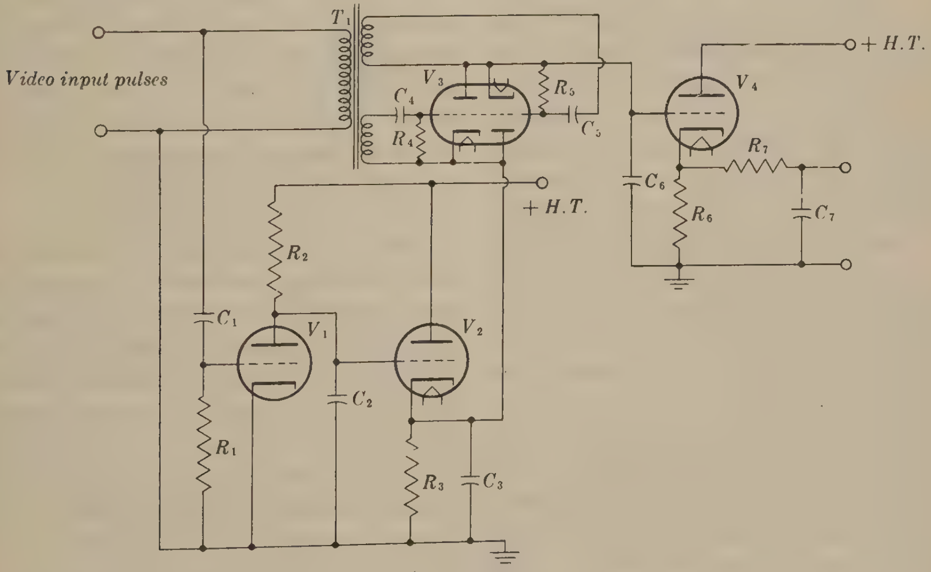


Fig. 13-45 Step-function demodulator for p-t-m receiver.

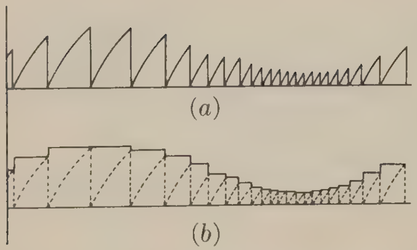


FIG. 13-46 Waveforms to illustrate operation of the step-function demodulator (a) Voltage across C_2 (b) Voltage across C_6.

than C_6 at the instant when the grids of V_3 are made positive, the anode of the right triode becomes positive to its cathode, the valve conducts and the voltage across C_6 rises until the capacitors have the same voltage. On the other hand, if C_3 is charged to a lower voltage than C_6 at the instant the grids are made positive, the cathode of the left triode becomes negative to its anode. This valve conducts until C_3 and C_6 have the same voltage. Only one side of the double triode may conduct at any one time. In this way, C_6 is allowed to *view* the peak sweep voltage for a brief time during the reception of each pulse. Since V_3 is instantly cut off again at the end of the pulse, C_6 maintains its maximum voltage during the interval between pulses. The waveform of voltage across C_6 is shown in Fig. 13-46. This voltage, when smoothed out by low-pass filters is the a-f output of the demodulator.

R_3 and C_3 in the cathode circuit of V_2 constitute a delaying circuit. In the absence of C_3, the cathode potential of V_2 would fall abruptly as C_2 discharged through V_1. This fall in potential would occur at the same time as the positive voltage applied to the grids of V_3. As a result, C_6 would *view* the discharge of C_2 instead of its peak voltage. To prevent this, C_3 is connected across R_3. When C_2 discharges abruptly through V_1, the grid potential of V_2 falls sharply. The cathode potential is prevented from falling as quickly due to the discharge of C_3 through R_3. Thus V_2 is cut off by the discharge of C_2 and the peak voltage across C_3 is maintained while the viewing action takes place. The charging time of C_3 is not delayed as this capacitor charges through the lower resistance of V_2 and discharges through R_3 as explained.

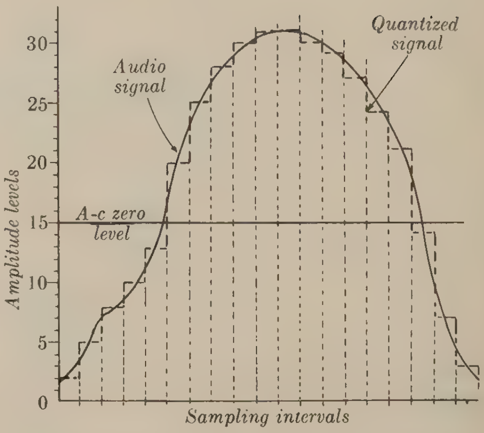


FIG. 13-47 Sampling and quantizing technique used in p-c-m.

The type of circuit just described is called a *step-function demodulator*. Such a circuit is preferable to the simpler one consisting of the sweep generator and low-pass filter because it makes possible the use of a lower pulse repetition rate for a given a-f modulating frequency without introducing undue distortion.

F. Pulse Code Modulation

Communication by pulse code modulation (p-c-m) is accomplished by *sampling* the audio signal at regular intervals and transmitting a coded group of pulses which represents the amplitude of the sample. Transmission of speech signals is considered to be satisfactory when the band-pass is extended to 3400 c/s which requires a sampling rate of about 8000 c/s. The sampling technique is illustrated in Fig. 13-47 which shows a portion of an audio signal sampled for

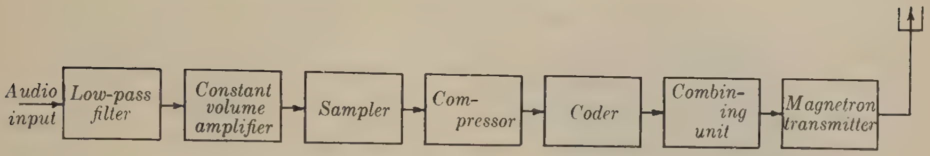


FIG.13-48 Block diagram of a p-c-m transmitter.

amplitude at twenty points during a period of $\frac{1}{400}$ second. In order to avoid the use of an excessively large number of code pulses to indicate the sample amplitudes, the signal is *quantized* into discrete steps or amplitude levels. The effective signal sent out from the transmitter and reconstructed at the receiver is indicated by the dotted outline of Fig. 13-47. Some distortion is therefore introduced by the quantization technique. The amount of distortion may be reduced by increasing the number of quantization levels but this requires the transmission of more pulses in each code group. As shown in the following paragraph, any amplitude level between 0 and 31 inclusive may be transmitted by a code group of 5 pulses. If the code group is enlarged to include 6 pulses, any amplitude level between 0 and 63 may be transmitted. It is found in practice that only 2% of distortion or *quantization noise* is introduced in transmitting speech signals with five-pulse code groups. In general then, code groups with more pulses may be used if greater fidelity is required; higher sampling rates may be used if higher audio frequencies are to be transmitted. Both of these extensions necessitate the transmission of more pulses per second and hence increase the bandwidth requirements.

1. BLOCK DIAGRAMS.

A block diagram of a p-c-m transmitting terminal is shown in Fig. 13-48. The audio signal is first passed through a low-pass filter which reduces the bandwidth to the 3400 c/s standard for speech signals. A constant volume amplifier is used to amplify the signal to a level at which the sampler operates most efficiently. The sampler effectively measures the amplitude of the signal at regular intervals 8000 times per second and quantizes each sample into one of 32 discrete amplitude levels. The compressor reduces the amplitude of each pulse by an amount proportional to its amplitude. This reduces the number of quantization levels without increasing the overall distortion. The coder produces a group of 5 pulses which, by their presence or absence only, indicate the amplitude of the sample. The code groups are then passed to a combining unit which combines the coded signals from other audio channels up to a maximum of 12 into one complex pulse train. This wave train which includes marker pulses for synchronizing the receiving equipment is used to modulate a magnetron transmitter which radiates pulses in the microwave band.

The receiving equipment is indicated by the block diagram of Fig. 13-49. The first unit sorts out the pulses into the

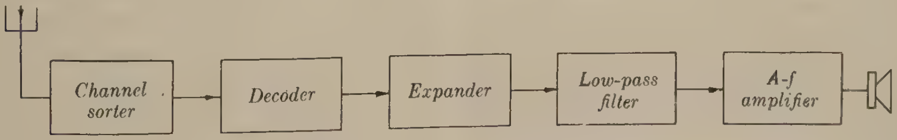


FIG.13-49 Block diagram of a p-c-m receiver.

12 channels combined at the transmitter. A decoder in each channel converts the code groups into pulses of varying amplitude, which, when expanded, are used to reconstruct the audio signal from the transmitter. A low-pass filter is employed to filter out the high-frequency components of the pulses produced by the decoding circuits. Suitable a-f amplifiers complete the receiving system.

2. BINARY NUMBER SYSTEM.

The *binary system* of numbers is useful for communication by p-c-m because the intelligence to be transmitted is conveyed only by the *presence* or *absence* of pulses at specific regular instants of time and not by their amplitude or waveform. This fact makes it possible to maintain communication even when the signal-to-noise ratio is comparatively high.

A comparison between the decimal and binary number systems is given in the table on this page. In the decimal system the columns reading from the left to the right side of a 5 digit group are ten-thousands, thousands, hundreds, tens and units. In the binary system, the corresponding columns of a 5 digit group are units, twos, fours, eights and sixteens. Thus the group 10011 in the decimal system is 10,000+10+1; in the binary system it is 1+0+0+8+16, or 25.

The binary number equivalents in electrical pulses are shown in the third column of the table. The binary number 1 is indicated by the presence of a pulse while the binary number 0 is indicated by the absence of a pulse. Hence, an amplitude level of 25 in the 5 pulse binary code would consist of a marker pulse followed by pulse, space, space, pulse, pulse.

Comparison of Decimal and Binary Numbers

DECIMAL NUMBERS	BINARY NUMBERS	ELECTRICAL CODE GROUPS
00001	10000	▮
00002	01000	▮
00003	11000	▮ ▮
00004	00100	▮
00005	10100	▮ ▮
00006	01100	▮ ▮
00007	11100	▮ ▮ ▮
00008	00010	▮
00009	10010	▮ ▮
00010	01010	▮ ▮
00011	11010	▮ ▮ ▮
00012	00110	▮ ▮
00013	10110	▮ ▮ ▮
00014	01110	▮ ▮ ▮
00015	11110	▮ ▮ ▮ ▮
00016	00001	▮
00017	10001	▮ ▮
00018	01001	▮ ▮
00019	11001	▮ ▮ ▮
00020	00101	▮ ▮
00021	10101	▮ ▮ ▮
00022	01101	▮ ▮ ▮
00023	11101	▮ ▮ ▮ ▮
00024	00011	▮ ▮
00025	10011	▮ ▮ ▮
00026	01011	▮ ▮ ▮
00027	11011	▮ ▮ ▮ ▮
00028	00111	▮ ▮ ▮
00029	10111	▮ ▮ ▮ ▮
00030	01111	▮ ▮ ▮ ▮
00031	11111	▮ ▮ ▮ ▮ ▮

3. SAMPLING CIRCUIT.

The sampling circuit (Fig. 13-48) receives the a-f input signal and produces a series of output pulses which are equally spaced and of constant duration. The amplitude of each output pulse is equal to the instantaneous amplitude of the input voltage. The operation of the sampler may be understood by referring to Fig. 13-50. The a-f signal is applied across the load resistor R_3 in series with two parallel triode circuits. One valve is connected to conduct during the positive half cycle of a-f input and the

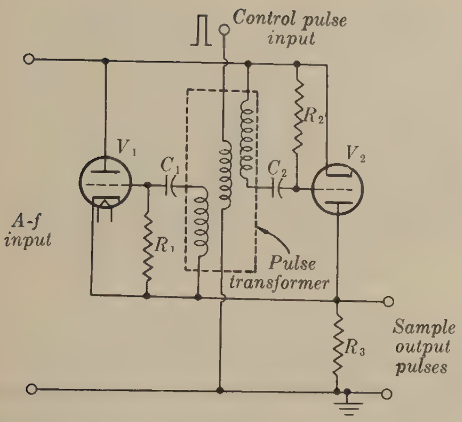


FIG.13-50 Triode sampling circuit

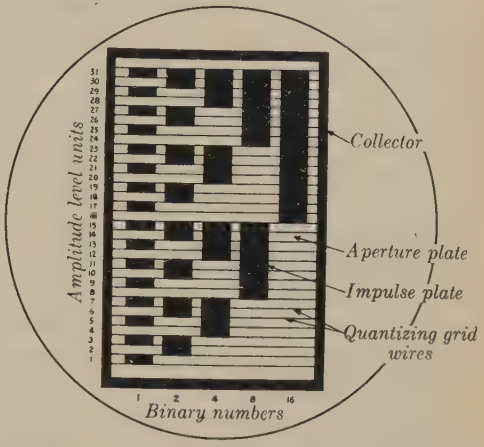


FIG.13-51 Diagrammatic view of coding device seen from gun end of tube.

other during the negative half cycle. Both valves are normally biased off by automatic bias developed by R_1C_1 and R_2C_2 in the grid circuits. The valves are permitted to conduct for a very short period once every $\frac{1}{8000}$ second by a positive pulse applied to both grids through the pulse transformer. If the a-f input is positive, electrons flow from ground through R_3 and V_1 while V_2 does not conduct. A positive pulse is developed across R_3 which is almost equal in amplitude to the signal at that instant. If the a-f input is negative, V_1 does not conduct and the electron path is through V_2 and R_3. A negative pulse is developed across R_3 equal to the amplitude of the input except for the small voltage drop across the valve.

4. CODER.

The sample pulses from the sampler are passed to the coding circuit (Fig. 13-48). This circuit translates each sample pulse into a coded group of five pulses which represents, in the binary number system, the amplitude of the sample. One type of coding device is illustrated in Fig. 13-51. It consists of a vacuum tube similar in construction to a cathode-ray tube with an electron gun and deflection plates. Mounted on the face of the tube is an *impulse plate* which receives the electron stream from the electron gun. Between the impulse plate and gun is an *aperture plate* which contains coded slots through which the electrons must pass to reach the impulse plate. Between the aperture plate and the gun is a *quantizing grid* framed in a rectangular metal *collector*.

The sample pulses are applied to the vertical deflection plates. The vertical deflection is held constant at a level determined by the pulse while the electron beam is swept horizontally across the aperture plate assembly. The beam is then cut off until the next pulse deflects it again. As the sweep takes place, the electrons pass through the apertures corresponding to that vertical level and are prevented from reaching the impulse plate where it is covered by the solid portions of the aperture plate. An electrical impulse is received from the impulse plate for each aperture that occurs in the path of the electron beam as the sweep takes place. The action is best understood by reference to a particular example. Suppose the sample pulse has an amplitude level of 25 which is 10011 in the binary number system.

This pulse causes the electron beam to be deflected vertically to the amplitude level indicated as 25 on the diagram of Fig. 13-51. As the horizontal sweep takes place from left to right the electrons pass through the apertures in columns 1, 4 and 5 but are obstructed by the aperture plate in columns 2 and 3. As a result, the impulse plate delivers a code group consisting of pulse, space, space, pulse, pulse. These pulses are amplified and are made to modulate the transmitters in the same sequence.

The horizontal quantizing grid wires and the collector ensure that the electron beam is swept across the plates at a definite quantizing level. If the beam falls on a wire, secondary emission takes place. The resulting electrons go to the collector and produce a bias which moves the beam into the space between the wires. A feedback amplifier circuit which provides a gain of 20 *db* provides the correcting bias.

5. Decoding Circuit.

The decoding circuit in the receiving equipment (Fig. 13-49) converts the groups of five code impulses from the transmitter into amplitude pulses from which the audio signal is reconstructed. The decoder consists of a resistor and capacitor in parallel, to which the signal is applied. The output is the voltage which exists across the capacitor at the end of the code group. This final voltage depends upon the number of pulses in the group, the time at which each is received and the amount of voltage lost by discharge through the resistor. The effect of each pulse on the capacitor voltage is illustrated in Fig. 13-52. Each of the 5 pulses received contributes the same amount of charge to the capacitor. The time constant of the *RC* combination is so chosen that the capacitor voltage falls to half its original voltage in a time equal to the interval between the pulses. Thus, if pulse 1 contributes 32 units of charge to the capacitor at t_1, it will have 16 units of charge at t_2, 8 units of charge at t_3, 4 units of charge at t_4 etc. At t_6 when the charge is measured, one unit of the original charge contributed by pulse 1 still remains. Similarly, pulse 2 which contributes 32 units of charge at t_2 will contribute 2 units of charge at t_6. The effect of each of the other pulses is shown in Fig. 13-52 as 4, 8 and 16 units respectively at t_6. The final result equals the sum of all the contributions received from the pulse groups. If all pulses are received, the final voltage is 1+2+4+8+16, or 31 units. If any pulse is missing its effect is missing from the final voltage. Thus the group 10011 referred to above will produce a final effect of 1+0+0+8+16, or 25 units. The volt-

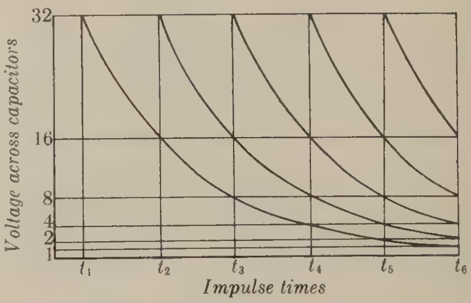


Fig.13-52 Operation of p-c-m decoder.

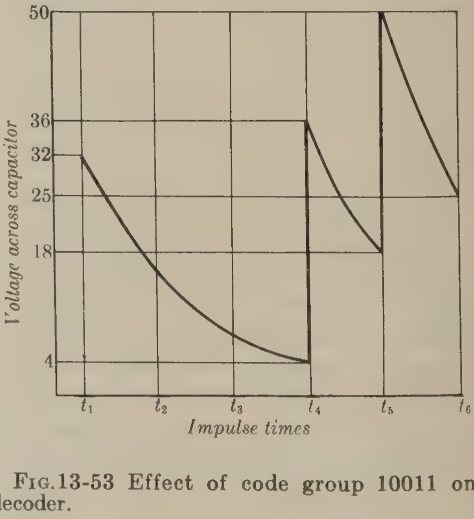


Fig.13-53 Effect of code group 10011 on decoder.

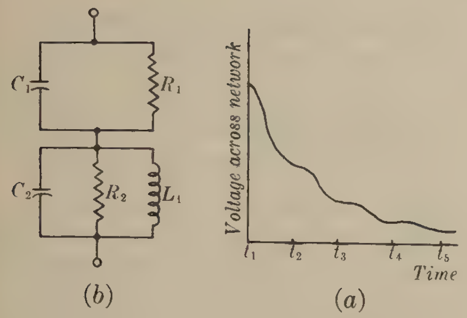


FIG. 13-54 Addition of *LC* circuit to decoder.

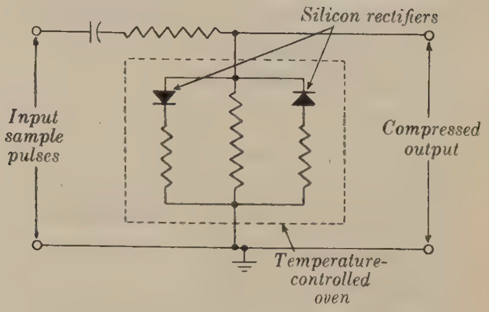


FIG. 13-55 Circuit of instantaneous compressor.

age changes across the capacitor produced by this group are shown in Fig. 13-53. The capacitor receives 32 units of charge at t_1. No pulses are received at t_2 and t_3 so that the effect of pulse 1 at t_4 is 4 units. At this time, pulse 4 contributes 32 units of charge making a total of 36. During the next interval, this charge falls to half its value, or 18 units. Another charge of 32 units is added at t_5 making a total of 50. During the next interval, this charge decreases to half its value or 25 units as indicated above.

The operation of the circuit may be made less critical by adding an oscillatory circuit in series with the R_1C_1 circuit as indicated in Fig. 13-54 (a). The period of L_1C_2 is also equal to the interval between pulses. The effect of L_1C_2 is to cause the exponential fall of voltage across the network to take the form of a series of steps as shown in Fig. 13-54 (b). In this way the output voltage is maintained nearly constant during the reception of each code impulse.

6. COMPRESSOR AND EXPANDER.

It has been pointed out that the quantization technique introduces some distortion which is referred to as quantization noise. The distortion can be reduced by increasing the number of quantizing levels but this requires the transmission of more pulses in each group. Since the ear has a logarithmic response to sound intensity, the quantizing voltage steps may be increased for signals of large amplitude and decreased for small amplitudes without loss of fidelity. A compressor circuit in the transmitting unit (Fig. 13-48) is used to perform this function. It causes the quantizing error to be greater when the signal amplitude increases, but maintains the fractional quantizing error constant for a wide range of amplitude variations.

The circuit of the instantaneous compressor is shown in Fig. 13-55. It is a voltage divider which includes two silicon rectifiers to provide non-linear resistance paths in parallel with the load circuit. The rectifiers have a short-circuiting effect on the signal which increases as the amplitude of the signal increases. One rectifier conducts for signals of one polarity and the other when the signal polarity reverses. Thus, steps of equal size after the compressor correspond to steps of graded size before the compressor.

The expander circuit in the receiving unit (Fig. 13-49) reverses the effect of the compressor. The expander is in fact an amplifier with a compressor circuit connected in its negative feedback path.

13.14 Questions and problems.

1. What is the chief disadvantage of amplitude modulation in a communication system?
2. How is this disadvantage overcome by frequency modulation?
3. Describe briefly how the capacitor microphone in Fig. 13-1 brings about modulation of the radio-frequency carrier.
4. What determines the frequency deviation of the modulated signal?
5. How may the frequency deviation of an f-m system be increased?
6. What circuit of an f-m receiver replaces the detector of an a-m receiver?
7. What is the purpose of the limiter stage?
8. Explain why a limiter is used in an f-m receiver and why it cannot be used in an a-m receiver.
9. Describe the operation of the detuned-circuit discriminator of Fig. 13-9.
10. What is the chief disadvantage of this discriminator?
11. What are the two main advantages of the coupled-circuit discriminator over the detuned-circuit discriminator?
12. State two advantages of the ratio detector over the discriminator.
13. How are these advantages achieved by the ratio detector?
14. What is the purpose of a reactance-tube circuit?
15. In what respect is the circuit of Fig. 13-18 an inductive circuit?
16. What is the purpose in having frequency multipliers following the reactance tube in f-m transmitters?
17. How is frequency drift prevented
    (a) in a-m transmitters
    (b) in f-m transmitters?
18. Outline briefly two distinct methods of providing frequency stability.
19. Why is it impractical to use f-m transmission at commercial broadcast frequencies?
20. Why is f-m communication practically limited to line-of-sight range?
21. What is meant by the term frequency spectrum as applied to a modulated signal?
22. What effect does an increase in amplitude of the modulating signal have on the bandwidth required for satisfactory transmission?
23. What effect does increasing the frequency of the modulating signal produce on the bandwidth requirements?
24. What is meant by the term deviation ratio?
25. How is the deviation ratio related to the number of sidebands in the signal?
26. What advantage does the phase-shift method of modulation offer in producing a frequency-modulated signal?
27. What amount of phase shift can be obtained by the modulator?
28. How is the phase deviation increased without increasing the frequency of the carrier unduly?
29. In what respect does the final p-m signal differ from the corresponding f-m signal?
30. Suggest a method of converting a p-m signal into an a-m signal.
31. State two advantages of transmission by the pulse-modulation method.
32. What determines the minimum number of pulses per second that must be transmitted to carry the modulation signal?

33. What is meant by the terms diplexing and multiplexing as applied to communication systems?
34. In the diagram of Fig. 13-37, what conditions determine the duration of the conducting period of the modulator valve?
35. What is the purpose of the synchronizing pulses transmitted by a multiplexing system?
36. What is the purpose of the gate valves in a pulse-width modulation receiver?
37. What features of multiplexing by p-w-m make it more applicable to telephone communication than to entertainment broadcasting?
38. What advantages does pulse-time modulation offer over other systems of modulation?
39. What is the purpose of the slicer circuit in a p-t-m receiver?
40. Explain the term step-function demodulator applied to the circuit of Fig. 13-45.
41. Explain the meaning of the terms sampling and quantizing as applied to pulse code modulation.
42. What factor determines the number of quantizing levels used in a given p-c-m system?
43. How must a p-c-m system designed for speech transmission be altered for use in transmitting higher frequencies without distortion?
44. Make a block diagram of a p-c-m transmitting terminal and state briefly the function of each unit.
45. Write the numbers 45, 51 and 63 in the binary system of numbers.
46. Draw a circuit diagram of a sampling circuit and explain its operation
    (a) when the input signal is positive
    (b) when the signal is negative.
47. What is the purpose of the aperture plate in the coding tube?
48. Explain the operation of the coding tube in coding a signal of amplitude level 13 units.
49. (a) Explain the operation of the decoding circuit when receiving a signal of amplitude 13 units.
    (b) State the voltage level of the capacitor when each pulse is received.
50. What is the purpose of the compressor unit of Fig. 13-48?
51. Draw a circuit diagram of a compressor and explain its operation for signals
    (a) of small amplitude
    (b) of large amplitude.

CHAPTER XIV

DIRECTION FINDING

Direction finding or d-f is the determination of the direction of an incoming radio wave. This is accomplished by the use of a directional receiving aerial and a sensitive receiver. Ordinary d-f employs the low and medium radio frequencies. At these frequencies the ground wave predominates and is vertically polarized (Chapter XII).

14.1 Loop aerials. The directional aerial which is the basic component in a d-f system is some form of *loop aerial.* Fig. 14-1 shows a simple loop aerial with its plane vertical and set at right angles to the line of propagation of the incoming radio wave. The wave front reaches the two vertical sides *A* and *B* simultaneously. As the wave passes through the loop, an alternating e.m.f. is induced in the two sides. These e.m.f.'s have the same amplitude at any instant but act in opposite directions around the loop. Hence the resultant e.m.f. across the ends *XY* is always zero. No voltage is induced in the horizontal sections of the loop since the lines of magnetic force are parallel to them (i.e. the lines of electric force are perpendicular to them). It is to be noted that a consideration of the electrostatic field of the wave and its effect on a conductor leads to the same conclusion as when the magnetic component is considered.

If the loop is rotated so that its plane lies along the line of propagation (Fig. 14-2) the wave front reaches side *A* before side *B*. As the wave passes, each of the vertical sides has an e.m.f. induced in it which depends upon its length and the strength of the passing wave at the instant. An alternating e.m.f. is developed around the loop. The sequence in

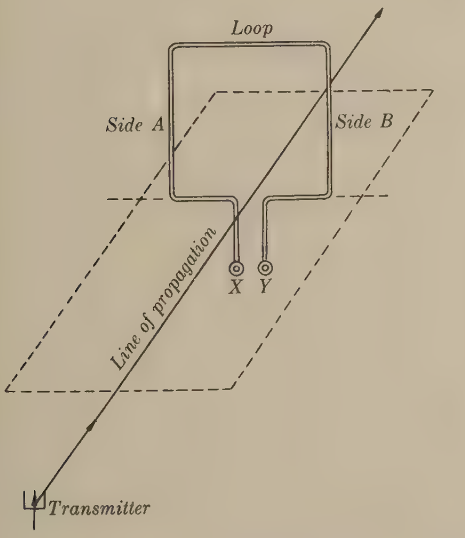


Fig.14-1 Simple loop aerial with its plane perpendicular to the line of propagation.

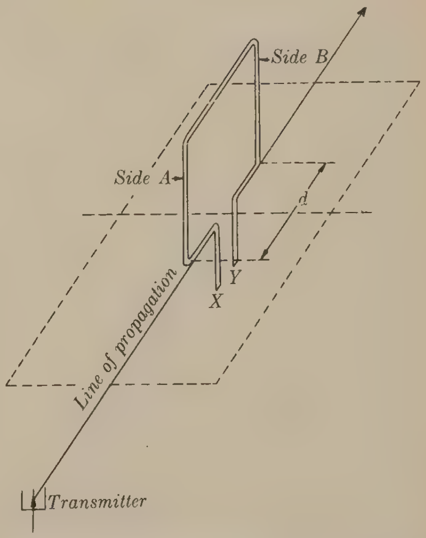


Fig.14-2 Simple loop aerial with its plane parallel to the line of propagation.

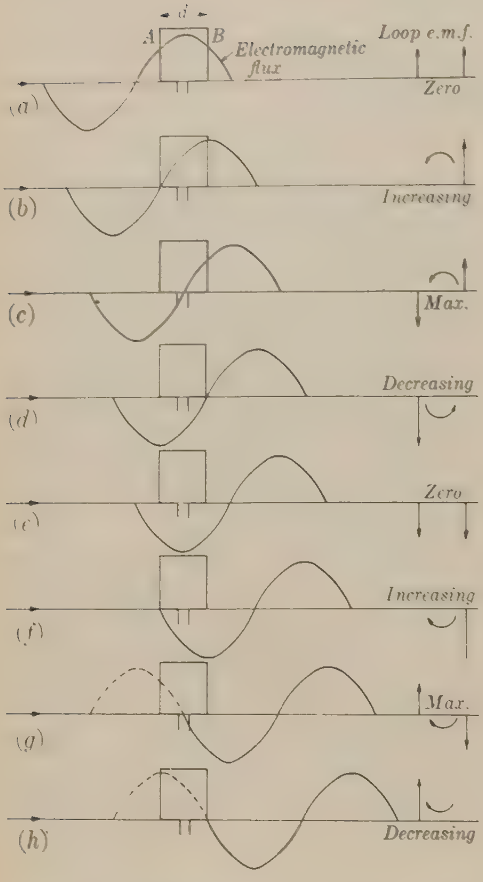


Fig. 14-3 Change in loop e.m.f. during one cycle of r-f wave.

the production of one cycle of the loop e.m.f. is shown in Fig. 14-3 (a) to (h). In Fig. 14-3 (a) the e.m.f.'s induced in the two sides are equal in amplitude but act in opposite directions around the loop. Hence the loop e.m.f. is zero. In (b) the e.m.f. in side *A* is zero while the e.m.f. in side *B* is a maximum. Hence there is a resulting e.m.f. across the loop. In (c) the e.m.f.'s in *A* and *B* act in the same direction around the loop and the resulting e.m.f. is a maximum. The remainder of the cycle is shown in Fig. 14-3 (d) to (h).

It is noted that zero loop e.m.f. occurs when the maximum field strength of the wave appears at the centre of the loop.

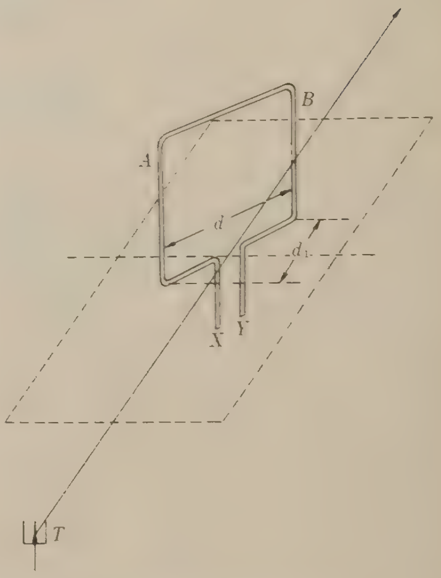


Fig. 14-4 Simple loop aerial with its plane making some angle between 0° and 90° with line of propagation.

Hence the loop e.m.f. is 90° out of phase with the alternating field flux. Maximum loop e.m.f. occurs when the voltages in the two vertical arms are equal and opposite. These voltages have their greatest values in opposite directions when the arms are $\frac{\lambda}{2}$ apart (Fig. 14-11). However, since modern receivers have high sensitivities, loop voltages obtained by using separations of a small fraction of $\frac{\lambda}{2}$ are satisfactory. The amplitude is increased by proper selection ot the length of the vertical arms and by the use of several turns for the loop.

If the plane of the loop is neither parallel to the direction of the wave nor perpendicular to it but occupies some intermediate position as in Fig. 14-4, the wave front reaches side *A* before side *B* but the effective width d_1 is less than *d* and therefore the amplitude of the loop e.m.f. is less than when the plane of the aerial is along the line of propagation. As the loop is rotated

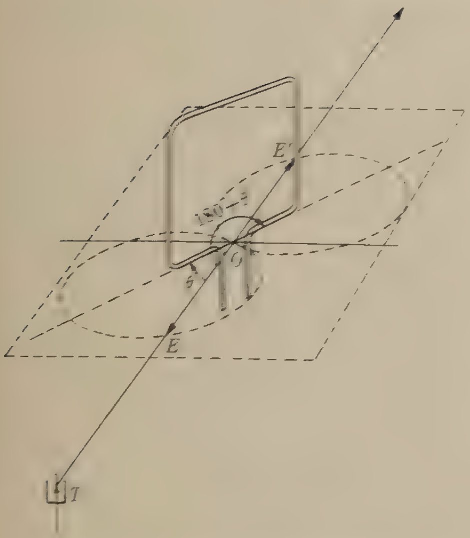


Fig.14-5 Polar diagram of loop aerial.

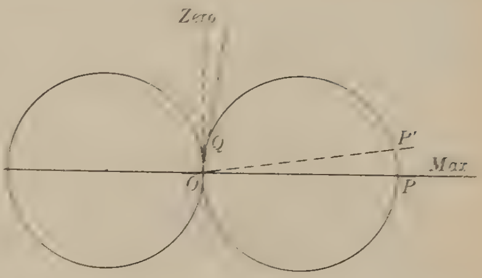


Fig.14-6 Comparison of change in loop e.m.f. for small change in bearing near zero and maximum positions.

through 360° the magnitude of the induced e.m.f. and hence the input to the receiver changes according to the angle which the plane of the loop makes with the line of propagation. The horizontal polar diagram of Fig. 14-5 shows how the amplitude of the e.m.f. changes as the loop rotates through 360°. This polar diagram forms a figure-of-eight pattern. Since the directional characteristics are an integral part of the loop aerial, the polar diagram is considered to rotate with the loop. Therefore *OE* represents the amplitude of the loop e.m.f. corresponding to the angle θ between the direction of the transmitter and the plane of the loop. Equal signals are received on bearings which are 180° apart. Hence *OE′* represents the amplitude of the loop e.m.f. for the angle $\theta+180°$. If θ is zero or 180° the loop e.m.f. becomes a maximum. That is, there are two positions 180° apart for zero e.m.f. and two positions 180° apart for maximum e.m.f. As a result the use of a rotatable loop aerial enables the line along which the transmitter lies to be found but does not give the sense; that is, it does not distinguish on which side of the receiver the transmitter is situated along this line.

14.2 Use of zeros for d-f. From Fig. 14-6 it may be seen that a small angular change in the vicinity of a maximum produces a very small change in signal strength as represented by *OP-OP′* However, in the vicinity of a zero, for the same angular change, the change in signal strength is relatively large as represented by *OQ*. Hence, to obtain a *d-f bearing*, the frame aerial is always rotated to the position of zero signal strength. A double-ended pointer placed at right angles to the plane of the frame then gives the *line of direction* of the transmitter but not its sense.

14.3 Determination of sense. To obtain the *sense* of the transmitter bearing, a combination of loop aerial and vertical aerial is used. The e.m.f. applied to the receiver from the vertical aerial is adjusted to be the same as the maximum value of the e.m.f. from the loop aerial. The circuit is designed so that during 180° of rotation of the loop aerial its e.m.f. is added to that of the vertical aerial and during the remaining 180° of the complete rotation its e.m.f. is subtracted from that of the vertical aerial. Fig. 14-7 shows the polar diagrams of the vertical aerial and the loop aerial together with the polar diagram of the combination which is obtained by adding the separate diagrams to the left of the d-f zero line and subtracting them

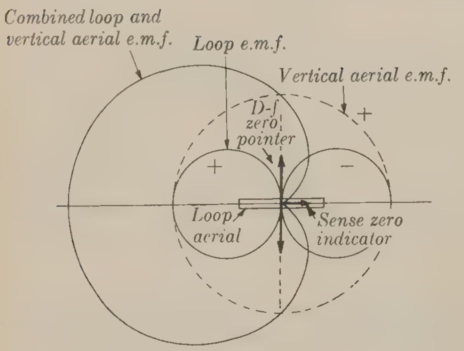
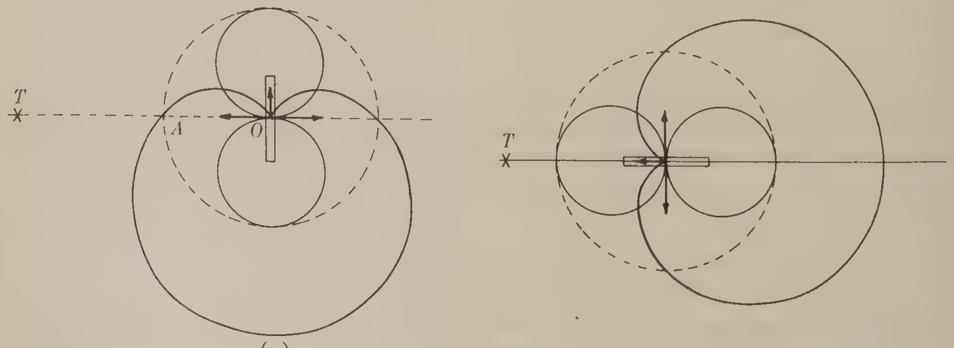


Fig.14-7 Polar diagram of combination of loop aerial and vertical aerial.

to the right. The polar diagram of the combination is a heart-shaped curve called a *cardioid*. As the loop is rotated with the vertical aerial switched in, there is only one position for zero receiver input and one for maximum receiver input. The position for zero receiver input is displaced 90° from the zero positions obtained with the loop aerial alone. The *sense indicator* is therefore at right angles to the *d-f pointer* (Fig. 14-7). Hence, after a zero position is found using the loop aerial alone, the vertical aerial is switched in and a signal of strength OA, Fig. 14-8 (a), is applied to the receiver. The loop aerial is then rotated through 90° to give a new zero signal, Fig. 14-8 (b). The *sense indicator* now indicates which of the two bearings obtained using the loop aerial alone is correct.

The cardioid diagram has a less distinct minimum than a figure-of-eight. That is, a small rotation of the loop when it is near the cardioid minimum does not cause as great a change in the intensity of input to the receiver as does an equal rotation near a minimum when the loop is used alone. Hence the combination is used only to indicate which of the two bearings obtained using the loop alone is the correct one. As a result, the cardioid minimum need not be exactly zero. The shape of the polar diagram obtained when the amplitude of the vertical aerial e.m.f. is greater than the maximum amplitude of the loop e.m.f. is shown in Fig. 14-9 (a). One minimum is obtained but it is more indistinct than for the perfect cardioid. Fig. 14-9 (b) shows the polar diagram when the vertical aerial e.m.f. is less than the loop e.m.f. Two minima are obtained with a small lobe of reception between. In practice, either of these conditions is acceptable if not too extreme, since only the sense is to be indicated.

14.4 Phase relationships in the determination of sense. It is seen in section 14.1 that a loop aerial gives the line of direction of a transmission only. That



(a) (b)

Fig.14-8 Polar diagram of combination for signal direction shown (a) when loop is in zero position (b) when combination is in zero position.

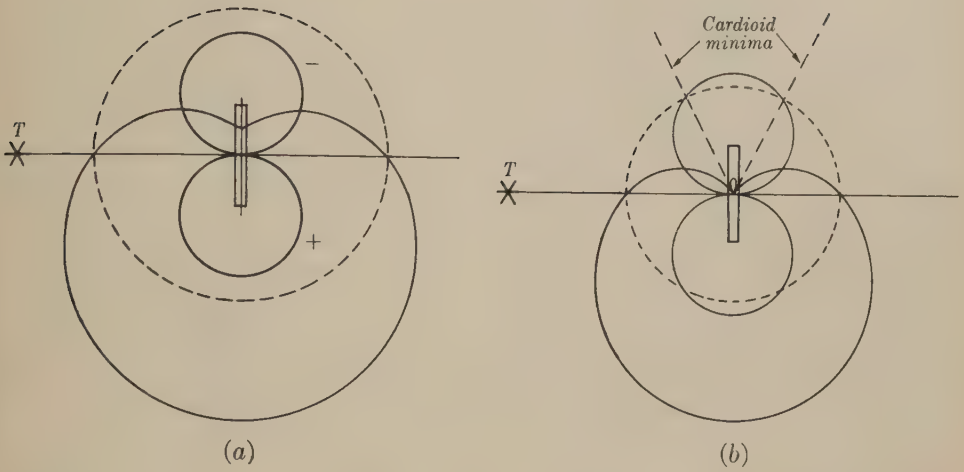


FIG.14-9 Polar diagram of combination of loop and vertical aerials when amplitude of vertical aerial e.m.f. is (a) greater than (b) less than maximum amplitude of loop e.m.f.

is, two bearings are determined, 180° apart, one of which is the correct bearing. To determine the correct bearing or sense of the transmission the output from a single vertical aerial mounted nearby is coupled to the receiver input along with the output of the loop aerial. It is pointed out also that the loop e.m.f. is 90° out of phase with the field flux of the wave at the centre of the loop. Since the loop dimensions are small compared to a wavelength (Fig. 14-10) the loop e.m.f. is practically 90° out of phase with the field flux at any position near the loop. Hence if a vertical aerial is erected near the loop, its e.m.f., being in phase with the wave, is also 90° out of phase with the loop e.m.f. The e.m.f. from the vertical aerial is coupled into the receiver input in such a way that a 90° phase

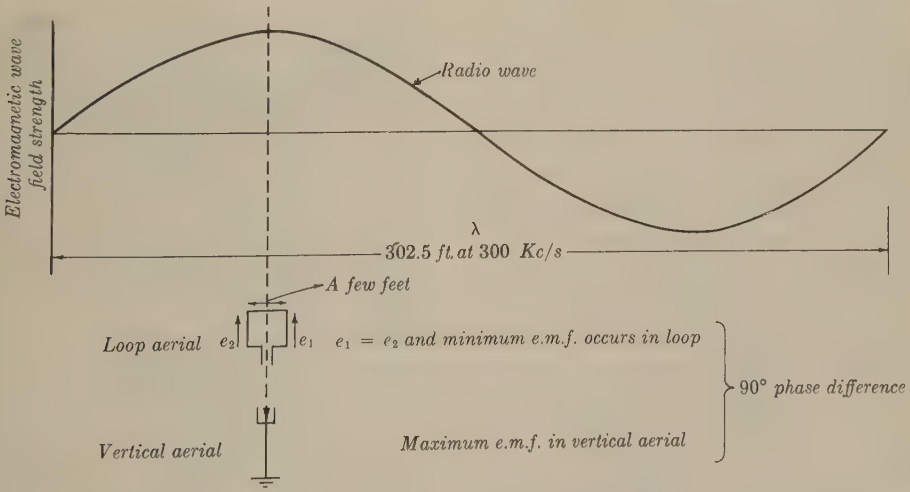

Electromagnetic wave field strength
Radio wave
λ
302.5 ft. at 300 Kc/s
A few feet
Loop aerial
e_2
e_1
$e_1 = e_2$ and minimum e.m.f. occurs in loop
90° phase difference
Vertical aerial
Maximum e.m.f. in vertical aerial

FIG.14-10 Phase relationship between loop e.m.f. and vertical aerial e.m.f. when loop width is small compared to a wavelength.

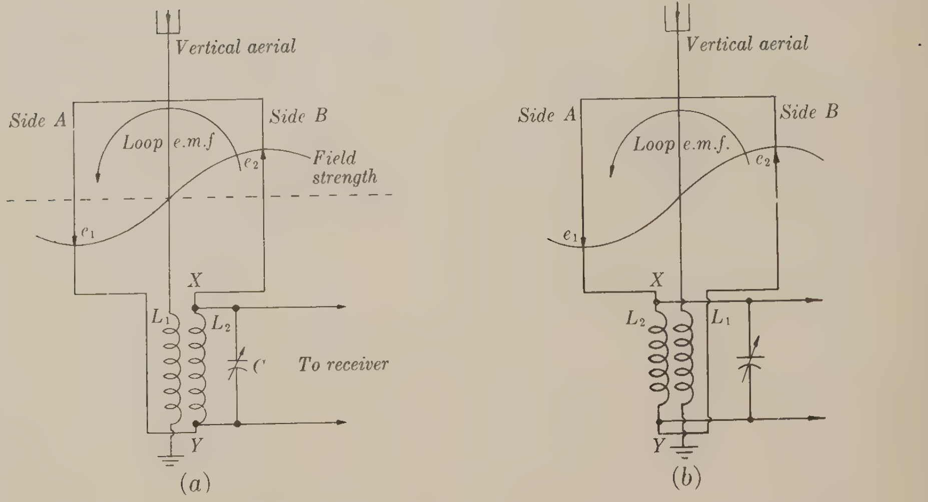


FIG. 14-11 Simple circuit to change phase of vertical aerial e.m.f. 90° relative to phase of loop e.m.f. with connections arranged for (a) zero receiver input (b) maximum receiver input.

shift occurs. Hence at the input to the receiver the e.m.f. from the vertical aerial is either in phase or 180° out of phase with the loop e.m.f. depending on the orientation of the loop. Therefore as the loop is rotated with the sense aerial in use, the vertical aerial e.m.f. will reinforce the loop e.m.f. for one of the maxima obtained with loop only but will weaken it for the other maximum 180° of rotation away. If the vertical aerial e.m.f. is adjusted to be equal to the maximum loop e.m.f., then complete cancellation takes place. Thus one of the maxima obtained in the receiver with the loop only becomes a zero and the other increases to double its value. A zero is thus obtained for only one bearing and the sense of the transmission is determined.

This method of sense determination depends on the 90° phase shift of the sense aerial e.m.f. relative to the loop e.m.f. A simple circuit to accomplish this is shown in Fig. 14-11. The vertical aerial is coupled to the loop output by the mutual inductance between L_1 and L_2. The vertical aerial circuit is made resistive so that the current in L_1 is in phase with the e.m.f. This may be done by tuning the vertical aerial to resonance by connecting a swamping resistance in series with L_1. The e.m.f. induced in L_2 by this current leads this current and hence the vertical aerial e.m.f. by 90° Thus the e.m.f. introduced into the receiver input from the vertical aerial is shifted 90° in phase so as to be either in phase or 180° out of phase with the loop e.m.f. The magnitude of the e.m.f. from the vertical aerial is varied by varying the mutual inductance between L_1 and L_2.

The phase relationship in the receiver input depends on the orientation of the loop. When the loop is rotated to a position of a maximum loop e.m.f. without the sense aerial, at some instant zero field strength occurs at the centre of the loop, Fig. 14-11 (a). The loop e.m.f. is then counterclockwise and a maximum at this instant. If a vertical aerial is placed at the centre of the loop, the e.m.f. induced in the vertical aerial is zero. The e.m.f. induced in L_2 leads this by 90° and therefore is a maximum and in the direction X to Y if the coils L_1 and L_2 are wound similarly. The

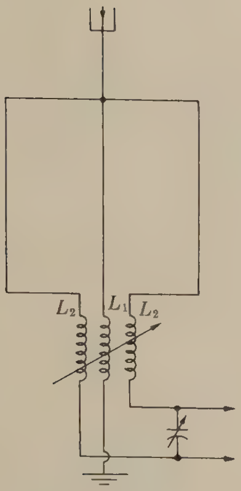


Fig.14-12 Circuit of Fig. 14-11 with L_2 divided into two sections for electrical balance.

loop e.m.f., being counterclockwise, is directed from Y to X so that the two voltages are 180° out of phase. If the e.m.f. due to the vertical aerial is equal to the loop e.m.f., they cancel each other and the input to the receiver is zero.

If, however, the loop were rotated through 180°, Fig. 14-11 (b), then the loop e.m.f. would be directed from X to Y and hence would be in phase with the e.m.f. in L_2 produced by the vertical aerial. The receiver input would then be twice the value of either voltage Therefore as the loop is rotated with the sense aerial being used, only one zero and one maximum is observed at the output of the receiver.

In practice, L_2 is divided into two sections (Fig. 14-12), one being placed in each side of the loop to maintain an electrical balance to offset vertical effect which is discussed in section 14.7.

14.5 The Bellini-Tosi fixed-frame aerial system. A rotating loop aerial may be used in shore installations but is not practical on board ship. Instead, a fixed-frame system known as the *Bellini-Tosi system* is used. It consists of two fixed loop or frame aerials with their planes vertical and bisecting at right angles (Fig. 14-33) mounted on the centre line of the ship. One loop aerial has its plane along the fore-and-aft line and the other athwartships. A device called a *radio goniometer* (Fig. 14-13) situated in the communications cabin is connected to the loop aerial. The goniometer consists of two coils also placed at right angles to each other as shown in Fig. 14-14. The 0°-180° coil

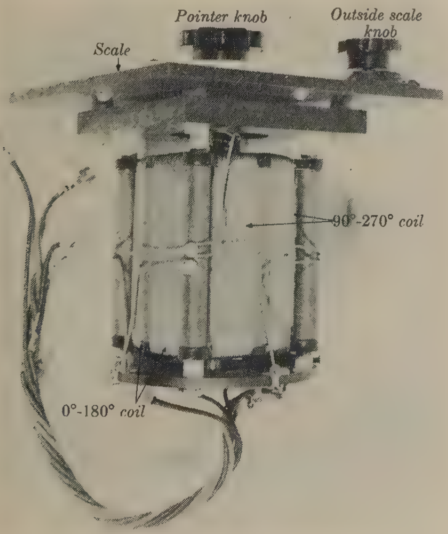


Fig.14-13 Radio goniometer.

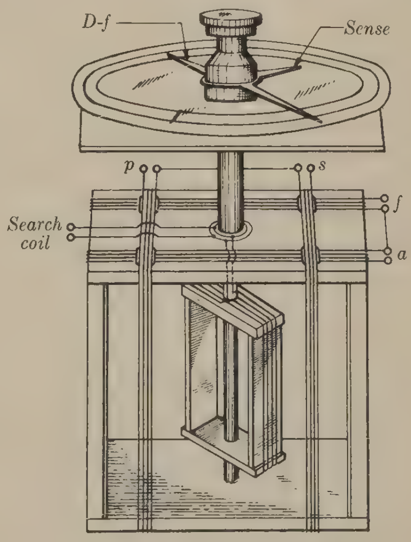


Fig.14-14 Line diagram of radio goniometer.

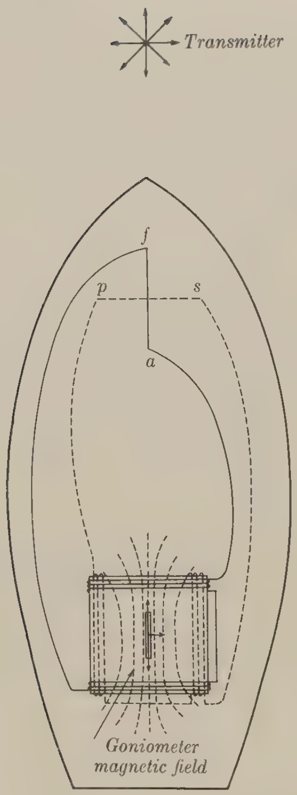


Fig.14-15 Goniometer field for signal with 0° bearing.

is connected to the port and starboard (p-s) loop and the 90°-270° coil is connected to the fore-and-aft (f-a) loop. A rotatable *search coil* is mounted at the centre of the crossed coils and is thus inductively coupled to them. The search coil is connected by means of slip rings to the receiver input. The operation of this system may be understood by examining the effects of signals with bearings of 0°, 90° and 45° in turn.

When the signal has a bearing of 0°, that is from dead ahead (Fig. 14-15), there is zero e.m.f. in the p-s loop. Thus there is no current in the 0°-180° coil of the goniometer and no magnetic field set up by this coil. There is, however, an e.m.f. in the f-a loop and hence a current flows in the 90°-270° degree coil. A magnetic field is set up inside this coil as shown in Fig. 14-15. Thus the magnetic field in the goniometer has the same direction relative to the goniometer 0°-180° line as the line of propagation has relative to the fore-and-aft line of the ship. If the search coil is rotated to make its plane parallel to both the direction of the magnetic field and the 0°-180° line, a minimum e.m.f. is induced in it and hence minimum receiver input is obtained. Thus the line-of-direction pointer is placed in the same line as the plane of the search coil. The line-of-direction pointer and scale are shown in the photograph of Fig. 14-16.

In order to obtain the correct sense of the bearing, the e.m.f. induced in a vertical aerial is coupled into the search coil circuit with the required 90° phase shift discussed in section 14.3. A 90° rotation of the search coil is then required to produce a new zero signal, Fig. 14-8 (a). The sense pointer fixed at right angles to the line-of-direction pointer then indicates the correct bearing.

If the signal is from 90° on the starboard beam (Fig. 14-17), the e.m.f. induced in the f-a loop is zero. Hence there is a current only in the p-s loop

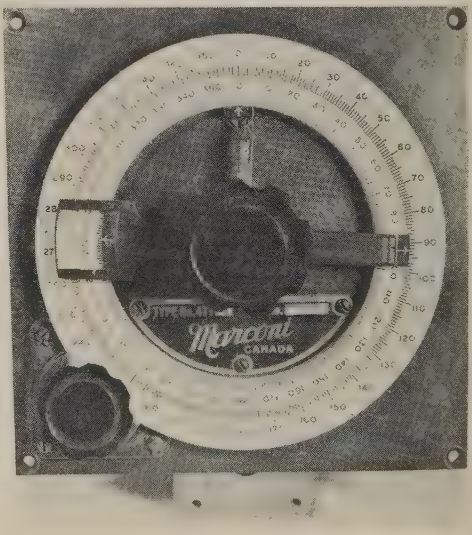


Fig.14-16 Goniometer scale and pointers.

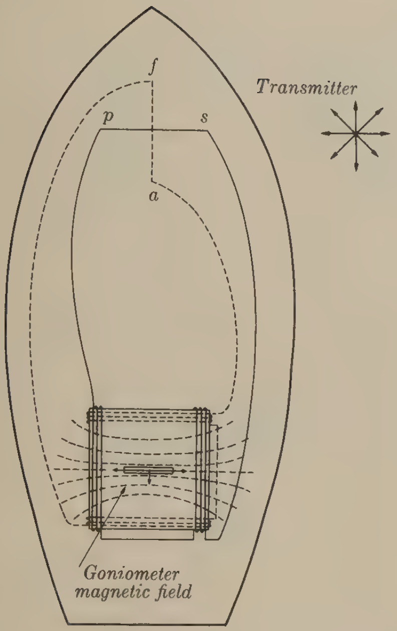


FIG. 14-17 Goniometer field for signal with bearing of 90°.

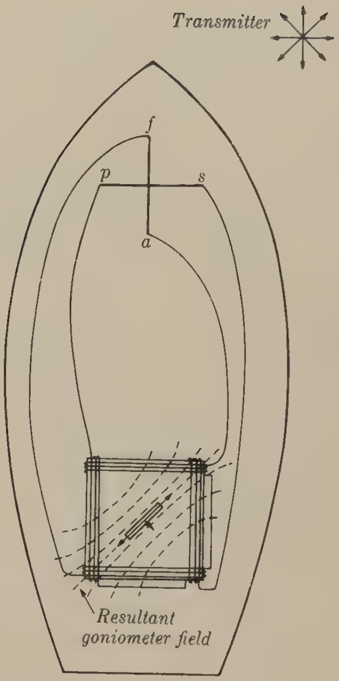


FIG. 14-18 Goniometer field for signal with a bearing between 0° and 90°.

and a magnetic field is set up in the goniometer parallel to the 90°-270° line. Thus the search coil must be rotated until its plane is parallel to this field, that is, in a 90°-270° direction to obtain a minimum signal. The line-of-direction pointer then indicates bearings of 90° and 270°. When the sense aerial is switched in, a rotation of 90° gives a new minimum and the sense pointer indicates the correct bearing.

If the signal is from some intermediate bearing, as in Fig. 14-18, e.m.f.'s are induced in both loops. For example, a signal with a bearing of 45° causes equal e.m.f.'s to be induced in the loops. Hence equal currents flow in the goniometer coils and the resultant magnetic field set up about the search coil is at an angle of 45° to the planes of the coils. For a minimum receiver input, the search coil must be rotated to be parallel to these lines. In this position the line-of-direction pointer indicates bearings of 45° and 225°. When the vertical aerial is switched in, a 90° rotation of the search coil to a new minimum gives the correct sense of the bearing.

14.6 Errors in direction finding. The bearings obtained by d-f systems are not always accurate because of blurred or displaced minima. Consequently the bearing obtained differs from the true bearing by a certain number of degrees. This variation is called the *error of the bearing* and may be caused by the following effects:

(i) aerial or vertical effect caused by an unbalanced loop-aerial circuit

(ii) re-radiation from the masts, hull and other ship superstructure

(iii) night effect caused by partial horizontal polarization of the incident radio wave

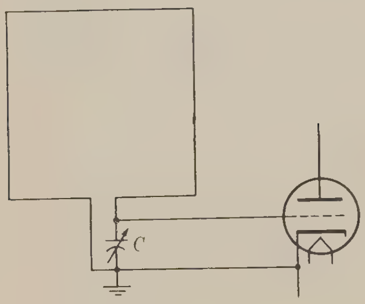


FIG.14-19 Simple loop-receiver input circuit.

(iv) aeroplane effect caused by horizontal polarization of the incident wave received from an aircraft due to the height of the aircraft or its use of a trailing aerial

(v) shore effect caused by the change in direction of a radio wave crossing from land to sea or vice versa.

The first two effects can be corrected and are discussed in the following sections, 14.7 and 14.8. The remaining errors, although they cannot be corrected, must be understood by the operator and are discussed in section 14.11. A loop which is properly designed to minimize the above errors and for which corrections are made may ordinarily give bearings which are correct to within $\frac{1}{2}$ to 2 degrees for nearby transmitters. The error increases, however, with greater distance between transmitter and loop and with higher frequency of transmission. The error is greater over land than water. The greatest distance for which satisfactory bearings are obtainable varies from 50 to 200 miles by day for frequencies about 500 Kc/s and up to several thousand miles by day at much lower r-f frequencies.

14.7 Vertical effect. Vertical effect occurs if the two vertical sides of a simple loop aerial are connected to ground by impedances which are unequal. This unbalance is caused by the actual circuits connecting the loop and the receiver and also by unwanted capacitive coupling introduced into the circuit by guy wires, gun mounts, masts and funnels.

Consider the simple loop circuit, shown in Fig. 14-19, with the voltage output of the loop, which is developed across the tuning capacitance C, applied to the receiver input. When the plane of the loop is perpendicular to the direction of the radio wave, equal but opposite e.m.f.'s should be set up in each vertical arm so that the resultant receiver input is zero. However, since the path to ground from each vertical arm is different and has a different impedance value, unequal currents flow in the arms and

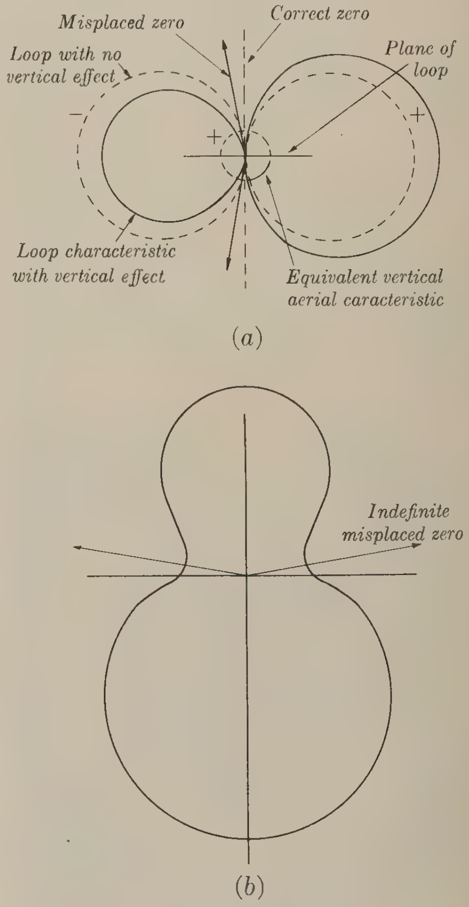

Misplaced zero
Correct zero
Loop with no vertical effect
Plane of loop
−
+
+
Loop characteristic with vertical effect
Equivalent vertical aerial caracteristic
(a)
Indefinite misplaced zero
(b)

FIG.14-20 Distortion of loop polar diagram due to vertical effect (a) Theoretical result (b) Practical result.

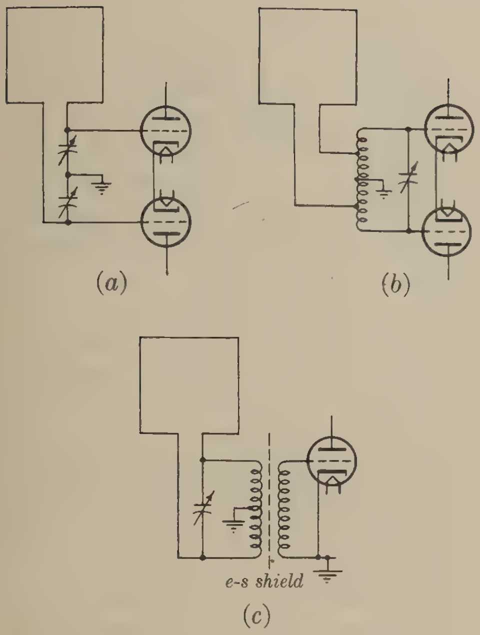


FIG. 14-21 Symmetrical loop-receiver input circuits (a) Push-pull input with split capacitor (b) Push-pull input with centre-tapped inductor (c) Input transformer with e-s shield.

hence the resultant current in the loop is not zero. Thus there is a small receiver input when there should be none. This effect is equivalent to that obtained by coupling into the loop circuit a small amount of the voltage from a nearby vertical aerial as this component is independent of loop orientation. Hence the figure-of-eight pattern is distorted as in Fig. 14-20, there being two zeros called *non-opposite minima* displaced from the correct zeros by a small angle. The presence of vertical effect can be observed by finding one zero and rotating the search coil through 180° and noting if another zero is obtained at that setting. If the zeros are not 180° apart, vertical effect is present and must be counteracted.

Vertical effect due to unbalance in the loop circuits is reduced by the use of a symmetrical loop circuit so that the two vertical arms have the same impedances to ground. The loop circuits may be made symmetrical by employing a push-pull input stage in the receiver as shown in Fig. 14-21 (a) and (b). Fig. 14-21 (c) shows another type of symmetrical loop circuit that does not use a push-pull stage.

Vertical effect due to unwanted capacitive coupling of guy wires and masts as well as unbalance in the loop circuits may be accurately eliminated by providing an additional voltage of opposite phase to cancel the unwanted voltage in the input circuit. By means of a small rotatable coil L (Fig. 14-22) coupled to L_1 and L_2, a small amount of voltage from a vertical aerial is introduced into the loop circuit. The vertical aerial must not be at resonance but reactive so that its current is 90° out of phase with the field. The voltage induced in L_1 and L_2 is 90° ahead or behind the loop e.m.f. depending on the orientation of L. During the first 180° of rotation of L the direction of its winding is the same as that of L_1 and L_2 and during the other 180° of rotation the direction of its winding is opposite. Thus the polarity and magnitude of the induced e.m.f. can be controlled by rotating L. Since the unbalanced voltage present due to ver-

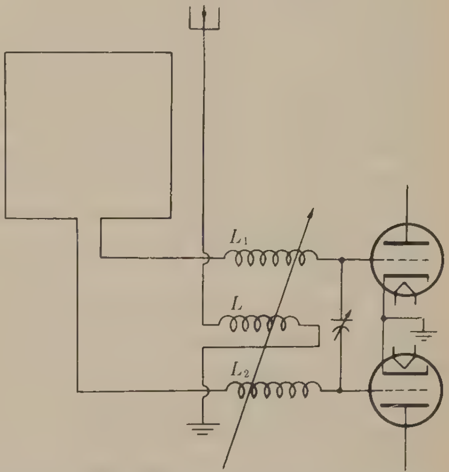


FIG. 14-22 Receiver input circuit to eliminate vertical effect, using a continuous control.

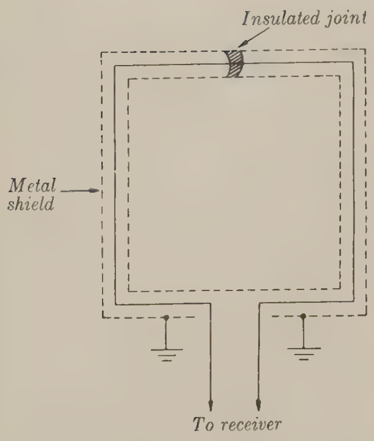


FIG.14-23 Grounded shield enclosing loop to reduce vertical effect.

tical effect is 90° in phase ahead or behind the loop e.m.f., it may be cancelled by the proper orientation of L. The proper setting of L is that for which two zeros are obtained exactly 180° apart.

Vertical effect due to unwanted capacitive coupling between the loop and nearby metallic objects is reduced by enclosing the turns of the loop in a grounded metal tube as in Fig. 14-23. The capacitance to ground of each vertical arm of the loop is equalized and made relatively independent of the loop position with respect to nearby metallic objects. The metal screen is prevented from acting as a loop aerial itself by having its two halves joined by an insulator at the middle of the upper horizontal arm. Hence no large circulating current can flow in the screen due to the e.m.f. induced by the impinging radio waves.

14.8 Errors in direction finding due to re-radiation. Indefinite and inaccurate zeros in d-f are also caused by re-radiation effects.

Near a shipboard d-f aerial there are many metallic masts and stays. The many electromagnetic fields which reach the ship from other transmitters and which are created by the ship's own transmitters cause e.m.f.'s of various frequencies to be induced in these metal conductors. Hence currents at these frequencies flow and re-radiation occurs. In the vicinity of the d-f loops there is, therefore, a complex electromagnetic field. At the frequency of the radiation on which a bearing is being taken, there is therefore the direct wave from the transmitter together with the many waves, smaller in amplitude, from the re-radiating structures. The resultant wave at this frequency will in general have a direction somewhat different from that of the transmitter. Thus the bearing observed using the d-f loops will be in error.

There are in general two kinds of ship re-radiators. The masts together with their stays and other vertical wires re-radiate as vertical aerials. The error caused by this re-radiation is known as *semi-circular error.* In addition there are loop re-radiators of which the most important is that made up by the two masts, the hull and the capacitance between the masts. This loop re-radiates as a loop aerial and since its component of re-radiation is about 90° different in phase from that radiated by the vertical wires, both radiation effects are different. The error produced in this case is called *quadrantal error.*

A. Semi-Circular Error

The resonant frequency of any of the vertical re-radiators is usually much lower than the frequency on which a bearing is being taken. Therefore these re-radiators are reactive so that their currents, and hence their radiations, are almost 90° out of phase with the e.m.f.'s induced in them and with the incoming wave. The re-radiation is thus in quadrature with the incoming radiation. When the search coil is set for a zero with respect to the incoming radiation

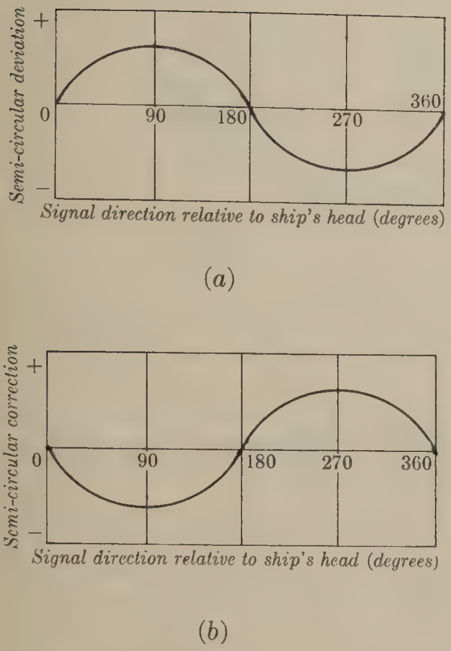


Fig. 14-24 Semi-circular error graphs to show (a) change of deviation with bearing (b) change of correction with bearing.

only, the quadrature radiation still causes a resultant e.m.f. to be set up and a zero is not obtained. Instead the search coil must be rotated to a new position for a zero thus giving an incorrect line-of-direction reading.

The d-f loops are usually situated on the central fore-and-aft line of the ship. The main vertical re-radiators are the masts usually aft of the loops. The direction of the re-radiation due to the vertical conductors is thus along the same fore-and-aft line of the ship. When the bearing of the incoming wave is 0° or 180°, therefore, the resultant wave has the same bearing and is merely changed in amplitude, so that the bearing error or deviation is zero. When the incoming wave has bearings of 90° or 270°, and the search coil oriented for a zero, it is in a position such that a maximum is obtained with respect to the re-radiated field. Hence maximum error occurs at these bearings. At intermediate bearings the error varies in magnitude between zero and the maximum. A graph of deviation against bearing is shown in Fig. 14-24 (a). Because the deviation changes sign after a half revolution of the bearing, this error is called *semi-circular error*.

The direction of the re-radiation field is always the same and its magnitude relative to that of the incoming field is constant. This means that the error caused by this re-radiation is always the same for any given bearing. Hence a *correction curve* is plotted, Fig. 14-24 (b), showing the amount and kind of the correction to be added to the observed bearing in order to get the correct bearing. Whether the correction is positive or negative in the first half of the curve depends on the position of the loop in relation to the masts, the main vertical re-radiators. For example, with the loop forward of both masts, the re-radiation has a 180° bearing. Hence all observed bearings between 0° and 180° will be too large (Fig. 14-25) and the correction is negative in the first two quadrants. Similarly all observed bearings between 180° and 360° are too small (Fig. 14-26) and require a positive correction.

Accurate corrections for bearings in all directions are made by means of a correction curve. This curve is plotted from observations of line-of-sight bearings and d-f bearings of a known transmitter both taken at 5° intervals as the ship is rotated through 360°. A correction curve of this kind is plotted when the ship is first fitted with d-f and at periodic intervals thereafter.

If the re-radiators are not symmetrically situated with respect to the loops the two bearings having zero error are not 0° and 180° but some other pair of opposite bearings. A correction curve can still be made up however.

The re-radiation has another effect

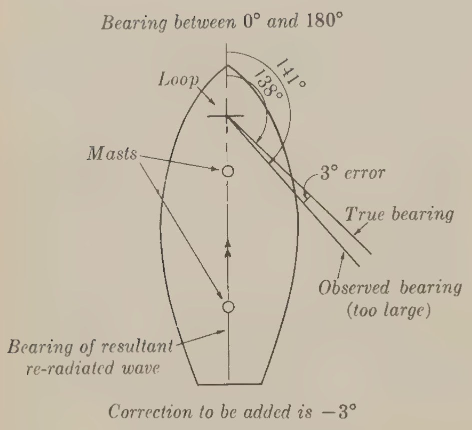

Bearing between 0° and 180°
Loop
141°
138°
Masts
3° error
True bearing
Observed bearing (too large)
Bearing of resultant re-radiated wave
Correction to be added is −3°

FIG. 14-25 Sign of semi-circular correction for bearings between 0° and 180° with loop situated forward of both masts.

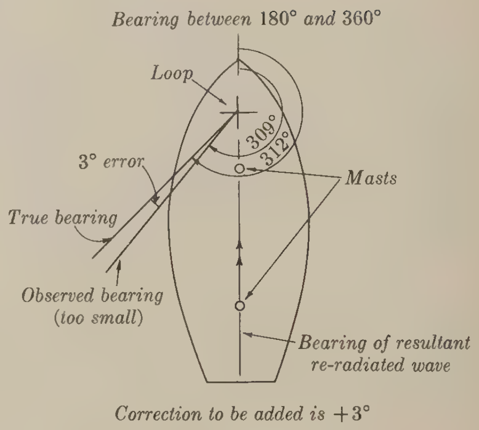
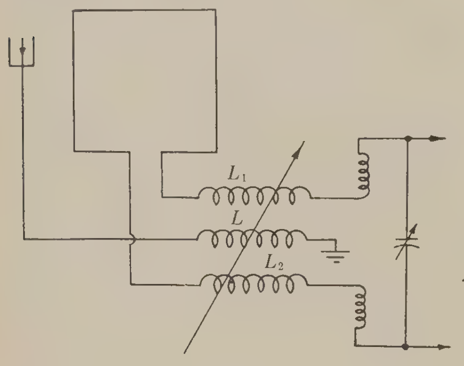
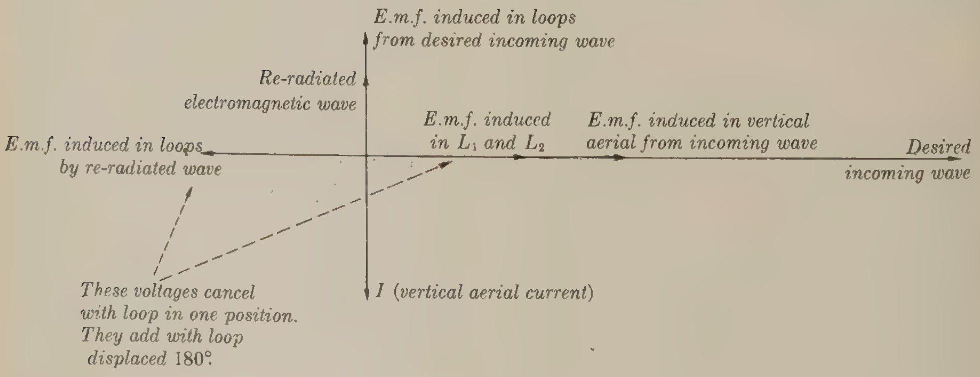

Bearing between 180° and 360°
Loop
3° error
309°
312°
Masts
True bearing
Observed bearing (too small)
Bearing of resultant re-radiated wave
Correction to be added is +3°

FIG. 14-26 Sign of semi-circular correction for bearings between 180° and 360° with loop situated forward of both masts.



(a)

besides that of causing zeros to be observed at incorrect bearings. The minima are also indefinite since the main re-radiation is in quadrature phase. Unless this is compensated for the correction curve is of little value. Fig. 14-27 (a) shows a method of producing a sharp zero and Fig. 14-27 (b) illustrates the phase relationships between the voltages involved. A small amount of voltage from a vertical aerial is introduced into the loop circuit by means of the coupling between L and L_1 and L_2. The vertical aerial must be operated so that it is not at its resonant frequency in order that


E.m.f. induced in loops from desired incoming wave
Re-radiated electromagnetic wave
E.m.f. induced in loops by re-radiated wave
E.m.f. induced in L_1 and L_2
E.m.f. induced in vertical aerial from incoming wave
Desired incoming wave
These voltages cancel with loop in one position. They add with loop displaced 180°.
I (vertical aerial current)

(b)

FIG. 14-27 Circuit to produce sharp zeros (a) Diagram (b) Phase relationships between voltages involved.

the current in it be 90° out of phase with the incoming wave. Since the e.m.f.'s induced in L_1 and L_2 by this current lead it by 90°, the e.m.f.'s induced in L_1 and L_2 are either in phase or out of phase with the wave. The unwanted e.m.f. set up in the loop by the re-radiation is 90° out of phase with the quadrature re-radiation and therefore in phase or out of phase with the incoming wave. Hence the correcting voltage is in phase or out of phase with the unwanted voltage so that if L is properly oriented relative to L_1 and L_2, the unwanted voltage is exactly cancelled by the correcting voltage. This correcting voltage is called a *zero-cleaning e.m.f.* This same voltage also acts to cancel vertical effect so that the same circuit eliminates both unwanted voltages to give a sharp zero. However, if the loop is rotated through 180° for the other zero the correcting and unwanted voltages are in phase so that the other zero is more indefinite or blurred than before. One sharp zero serves, however, to give a sufficiently accurate line-of-direction reading.

In most d-f systems, a switching arrangement is used so that one vertical aerial may be used both for sense determination and for balancing. For the sense determination the aerial must be operated at resonance so that the voltage introduced into the loop circuit is in phase with the loop voltage. For balancing, the aerial must be reactive in order that the voltage introduced into the loop circuit is 90° out of phase with the loop voltage and hence either in phase or 180° out of phase with the unwanted e.m.f.

B. Quadrantal Error

An additional re-radiation field is caused by the loop circuit composed of the two masts, the ship's hull and the capacitance between the masts. This loop is in the fore-and-aft plane of the ship so that its field induces no e.m.f. in the p-s loop but affects only the f-a loop of the Bellini-Tosi system. When the signal has a bearing of 0° or 180° this loop receives and re-radiates in the same line of direction so that no bearing error results. When the signal has a bearing of 90° or 270° the plane of the re-radiating loop is perpendicular to the incoming wave so that the e.m.f. induced in it is zero and there is no re-radiation. Hence the bearing error due to the effect of this loop is zero for these bearings also. The error is a maximum thus for bearings of 45°, 135°, 225° and 315°. A graph of the error against bearing is obtained using transmissions of known bearing and appears as in Fig. 14-28 (a). Since the error changes in each quadrant from zero to a maximum and back to zero it is called *quadrantal error.* It is usually much larger than semi-circular error aboard ship. As in the case of semi-circular error a correction curve can be plotted, Fig. 14-28 (b), showing the number of degrees to be added at any observed bearing to give the correct bearing.

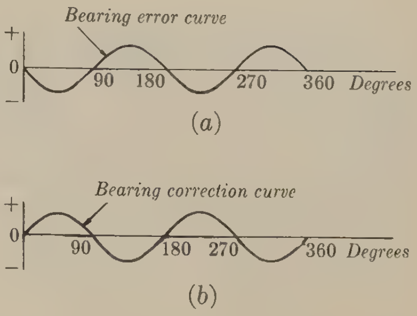


FIG. 14-28 Quadrantal error graphs (a) Bearing error curve (b) Bearing correction curve.

The correction to be added may be plus or minus in the first quadrant depending on the position of the loop relative to the mast-hull loop. If the d-f loops are situated forward of the mast-hull loop the correction is positive in the first quadrant, as illustrated in Fig. 14-

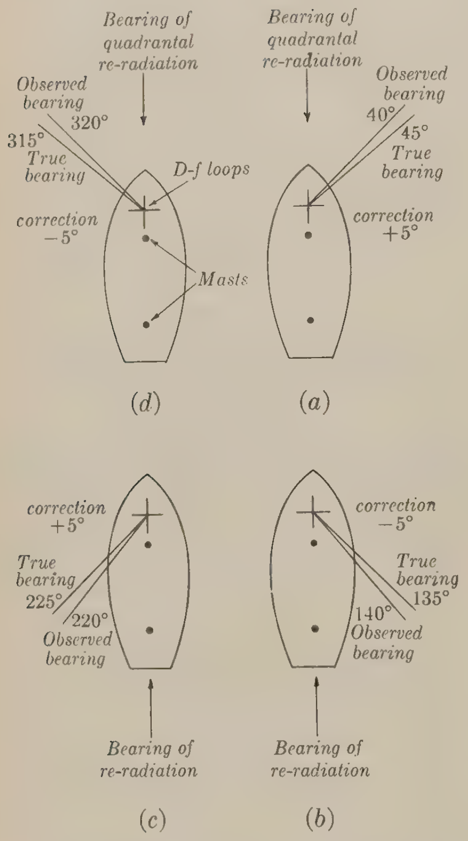


Fig.14-29 Sign of quadrantal correction with loops forward of masts for bearings between (a) 0° and 90° (b) 90° and 180° (c) 180° and 270° (d) 270° and 360°.

29 (a). The correction is minus in the second quadrant, Fig. 14-29 (b); positive in the third, Fig. 14-29 (c), and negative in the fourth, Fig. 14-29 (d). To avoid the use of a correction curve the pointer is often attached to the shaft of the rotary search coil via a cam arrangement. The cam is shaped according to the correction curve so that the pointer always indicates the true bearing.

The effect of quadrantal re-radiation is to produce a fore-and-aft component of radiation for any bearing except the cardinal bearings. The effect is counteracted in low-frequency d-f systems by making the f-a loop smaller than the p-s loop so that the total e.m.f. induced in the f-a loop by the direct field and the re-radiation field equals the e.m.f. induced in the p-s loop by the direct field alone. The two loop voltages are equal only at one frequency however. The size of the f-a loop is usually chosen to equalize these voltages at the lowest frequency in the range over which the d-f system is operated.

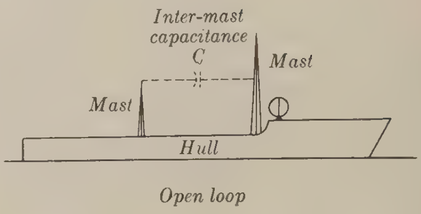


Fig.14-30 Mast-hull loop.

In addition to the adjustment in loop size a circuit adjustment is often made. A variable inductance called a *calibration choke* or *corrector* is connected in parallel with the 0°-180° coil of the radio goniometer. The value of this inductance is adjusted until the current in the 0°-180° coil equals that in the 90°-270° coil. This adjustment is complete at only one frequency since the reactance of the choke varies with frequency. To facilitate re-adjustment of this variable inductance it is provided with taps. The positions of the taps are determined during the initial calibration of the equipment, there being one tap for each frequency. In operation then the operator merely switches to the correct tap for any given frequency.

Quadrantal error may also be reduced at low frequencies by rigging a compensating *re-radiation loop*. The re-radiating loop formed by the hull and the two main masts is an open radiator since the circuit is completed by the small capacitance between the masts (Fig. 14-30). At low frequencies, such as those in the range 500 Kc/s to 1500 Kc/s, the resonant frequency of this loop is higher than any of these frequencies. Hence

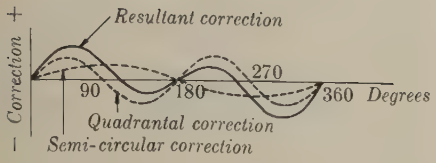


Fig. 14-31 Single curve to show total correction due to semi-circular and quadrantal corrections.

this open loop, consisting of inductance, capacitance and resistance in series, is capacitive. The compensating loop consists of three horizontal wires strung several feet above the d-f loop between the masts. This closed loop has a resonant frequency much lower than the range of frequencies being received and is inductive. Hence the currents flowing in these two loops induced by the incoming wave are 180° out of phase. The re-radiated fields are therefore also 180° out of phase and tend to cancel each other. Thus the quadrantal error is reduced. A compensating loop is required on ships where the quadrantal error is undesirably large, for example, of the order of 15° or more.

When the d-f equipment is first calibrated the correction curve obtained is a combination of the semi-circular and quadrantal correction curves since, in general, both effects are present. Such a combination curve is shown in Fig. 14-31. In such a case the deviation curve will have a similar shape.

14.9 Practical examples of low-frequency d-f aerial circuits.

A. *Example 1*

Fig. 14-32 shows the aerial circuit of a low-frequency d-f set using the Bellini-Tosi system. The aerial (Fig. 14-33) is a pair of fixed loops mounted on the centre line of the ship. Each loop is circular, the f-a loop having a slightly smaller diameter than the p-s loop to reduce quadrantal effect. Also each loop consists of several turns of copper wire surrounded by a grounded shield of metal tubing to reduce vertical effect. The shields are electrostatic shields only, since they are broken at the top by insulating joins. A long-wire vertical aerial is strung nearby to serve as a sense aerial. The fixed aerial loops are connected to the corresponding field coils of the goniometer which is similar to that shown in Fig. 14-14. In parallel with the f-a loop is the calibration choke L_6. The taps on L_6 are adjusted during the initial calibration and periodic checkups of the equipment to produce as small an error as possible on a bearing of 45°, at a frequency in the middle of the range of the equipment. The receiving powers of the two loops are equalized by this adjustment. The search coil is connected by means of slip rings and wire brushes to the search coil circuit. C is the basic fixed tuning capacitance and C_1 the variable tuning capacitance for the search coil circuit. C_1 is varied by means of the search tuning control on the front panel. C_2 is a trimmer capacitance used to align the search coil circuit so that the receiver tuned circuits and the search coil circuit are tuned to the same frequency when their dials show the same reading. C_3 is one section of the receiver tuning capacitance.

S_1 is the D-F—SENSE switch. In the d-f position the vertical aerial is disconnected as a sense aerial so that a line-of-direction reading may be made. However the vertical aerial is still in the circuit to supply a balancing e.m.f. for eliminating the effects caused by circuit unbalance and quadrature re-radiation. It is connected to ground through L_3 and L_4 in parallel with L_5. These inductances make the vertical aerial circuit reactive so that the aerial current induces an e.m.f. into the search coil circuit of the correct phase (Section 14.7). This balancing e.m.f. is injected into the search coil circuit by the variable mutual inductive coupling between L_3, L_4 and

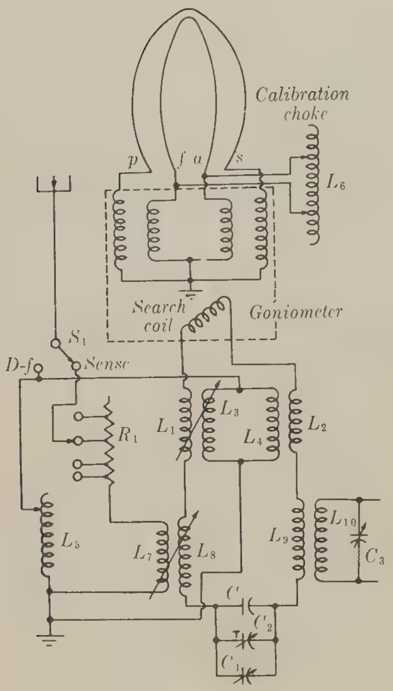


FIG.14-32 Low-frequency d-f aerial circuit.

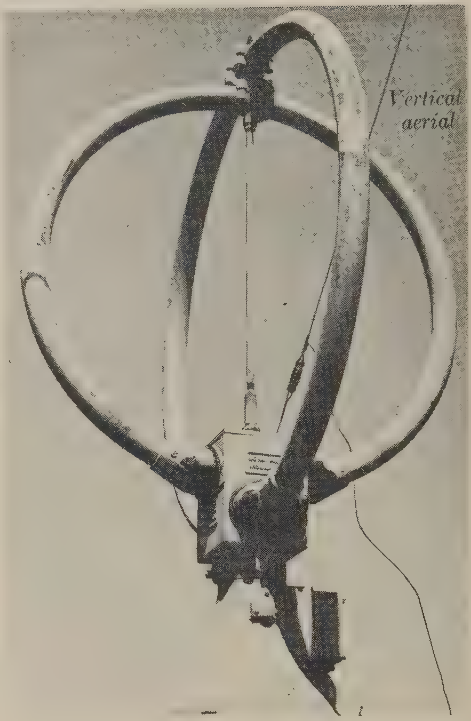


FIG.14-33 Fixed-frame aerial used with circuit of Fig. 14-32.

L_1, L_2. L_5 is adjusted during calibration or checkup to give an approximate balance and is called the *loading inductance*. The coupling between L_3, L_4 and L_1, L_2 is variable through 360° so that the polarity of the e.m.f. induced may be reversed and its amplitude controlled. This more accurate coupling control is adjusted by a knob on the front panel labelled BALANCE. When this knob is set at zero the coupling is zero. Rotation in one direction increases the amplitude of the coupled voltage with one polarity, and in the other direction increases the amplitude but with the opposite polarity.

In the SENSE position of S_1, the vertical aerial is connected to ground through R_1 and L_7. R_1 is a swamping resistance the value of which is variable by means of taps. The value of R_1 is chosen to be large compared with the aerial reactance so that the vertical aerial is almost exactly resistive. Hence the aerial current is in phase with the field, this condition being required for sense determination as discussed in section 14.4. The e.m.f. induced in the aerial is coupled into the search coil circuit by means of the mutual inductive coupling between L_7 and L_8. The amount of this coupling is adjusted so that the sense aerial input is sufficient to give a sharp zero for sense determination. The resultant search coil voltage is coupled to the receiver input by means of the mutual inductive coupling between L_9 and L_{10}.

The d-f receiver is usually a superheterodyne receiver of high sensitivity. The r-f and i-f circuits are carefully shielded so that no direct pick-up of signals occurs. Direct pick-up would not, in general, have the correct phase and would produce indistinct and displaced minima.

1. General Procedure in Taking a L-F D-F Bearing Using the Aerial Circuit of A.

The following listed summary illustrates normal procedure in operating the d-f equipment described in *A*. The ship should be held on a steady course while the bearing is being taken.

(i) Switch on receiver.

(ii) Place D-F—Sense switch to sense for searching for and tuning in the signal of which the bearing is required. Turn the C-w switch on during this step. Operate the main tuning and search tuning controls at the same time. The main tuning is used first. Then adjust the search tuning for maximum signal.

(iii) Switch off b.f.o. if m-c-w or r-t is being read.

(iv) Switch to D-F.

(v) Set balance control to zero and begin taking bearings by rotating the search coil from 0° clockwise through a complete revolution. Two maxima and two minima are observed. Rotate search coil again and adjust its position to one of the minima, reducing the volume so that the minimum or zero can be readily distinguished. Adjust the balance control to give a clearly defined zero. Then increase the volume slightly and re-balance until the zero is more clearly defined. The zero should be present over not more than 1° of rotation. This accuracy gives what is known as a first class bearing. Take the reading of the pointer on the relative bearing scale. Turn the balance control to an equal reading on the opposite side of the zero reading on its dial and then rotate the search coil through 180° to read the bearing for the other zero.

(vi) Switch to Sense. Rotate the search coil through 90° to obtain a new zero. The sense pointer now indicates the correct bearing of the two bearings obtained in step (v).

B. Example 2

Fig. 14-34 shows the circuit diagram of another d-f aerial circuit which uses some different devices. It can be operated over the same frequency range but is more flexible in its application. A photograph of the aerial is shown in Fig. 14-35.

A Bellini-Tosi system is used, consisting of the two fixed loops, the field coils L_1 and L_2 and the search coil L_3 similar to that of examble 1. L_9, in parallel with the f-a coil is the calibration choke adjusted as described in example 1. The switch S_1, which has four positions, grounds the f-a loop when turned to position 2; when turned to position 3 it grounds the p-s loop. By alternately grounding one loop and then the other, the two loop currents can be made equal for a given signal by the proper setting of the tapping point on L_9. The setting of the tap is controlled by a front panel adjustment for convenience. The correct setting for each frequency range is determined during calibration. For regular operation switch S_1 is used in position 4.

The aerial switch S_2 is a four-position, three-pole switch. The positions are labelled Search, Loops, Corrector and Sense on the front panels. Fig. 14-36 and Fig. 14-37 show the circuit for each position.

In the Search position 1, Fig. 14-36 (a), the vertical aerial is connected directly to the grid of the first r-f stage in the receiver through C_1, C_5 and C_6. This directly-coupled vertical-aerial voltage swamps the directional voltages of the loops and provides a large non-directional signal so that the receiver may be accurately tuned to the transmission whose bearing is required. The switch S_1 may be turned to its position 1

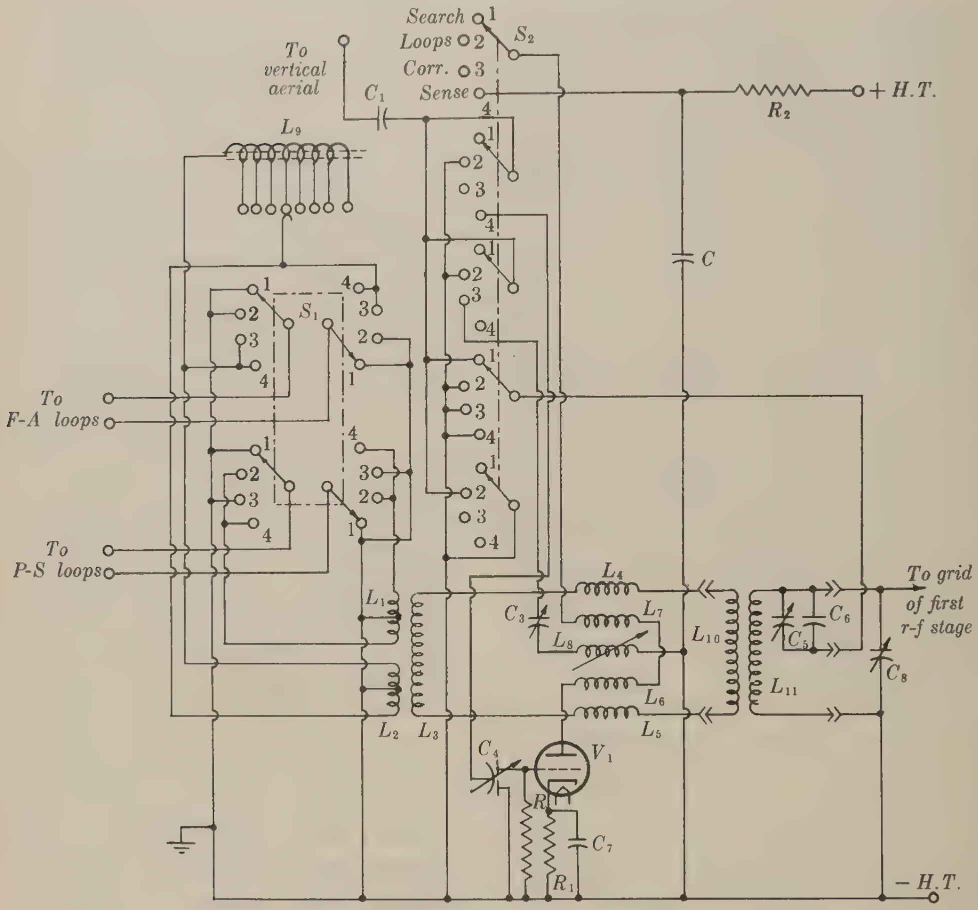


FIG. 14-34 Another l-f d-f aerial circuit.

to ground both loops if necessary for this preliminary tuning.

In the LOOPS position 2, Fig. 14-36 (b), the vertical aerial input voltage is grounded so that the only input to the receiver is from the two fixed loop aerials via the goniometer. The switch S_1 must be in position 4 so that both fixed loops are connected to the goniometer circuits. The line of direction of the transmitter is now found by rotation of the search coil. In this position of the aerial switch S_2, and in positions 3 and 4, C_5 is connected to act as a padder capacitance and C_6 as a trimmer capacitance for the first r-f amplifier tuned circuit.

In the CORRECTOR position 3, Fig. 14-36 (c), the vertical aerial is connected

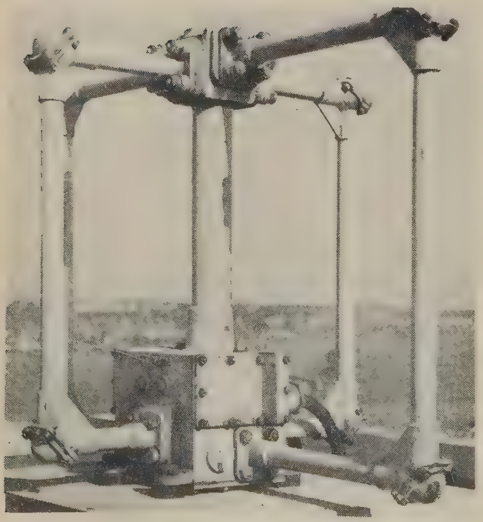


FIG. 14-35 Fixed-frame aerial used with circuit of Fig. 14-34.

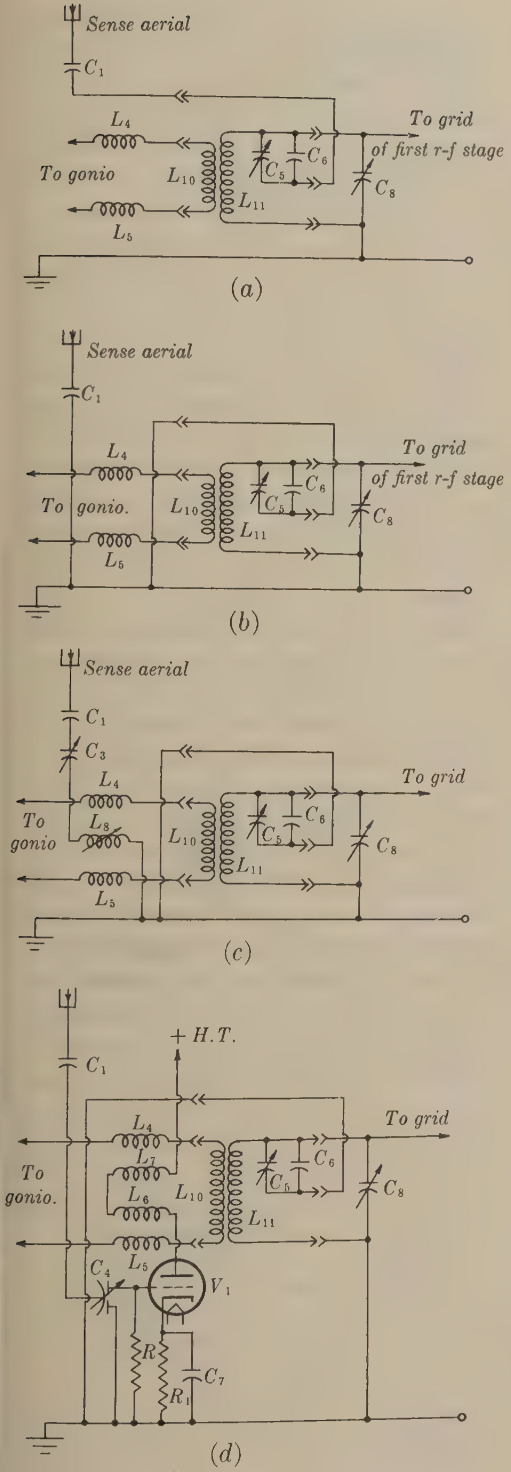


Fig.14-36 Circuit for each position of switch S_2 of Fig. 14-34 (a) Search position (b) Loops position (c) Corr. position (d) Sense position.

through C_1, C_3 and L_8 to ground. The vertical aerial is made inductive by L_6. A small amount of the voltage induced in the vertical aerial is coupled into the search coil circuit to act as a balancing voltage by the mutual inductive coupling between L_8 and L_4 and L_5. This coupling is variable so that the amplitude and polarity of the balancing voltage may be varied and is controlled from the front panel under the label Semi-Circular Corrector. This control must be adjusted for each frequency and for each new bearing. The phase of the balancing voltage may also be varied over a small range by means of C_3. This is a preset control and is adjusted for as small as possible resultant bearing deviation resulting from the use of the balancing e.m.f.

In the Sense position 4, Fig. 14-36 (d), the vertical aerial voltage is introduced to the search coil circuit by means of the valve V_1. This method of coupling has the advantage of producing a more accurate phase difference of 0° or 180° between the vertical aerial voltage and the resultant loop voltage. The grid-leak resistance R swamps the reactance of the aerial to make it resistive. The current through V_1 which is in phase with the voltage across R is in phase with the aerial current and voltage and hence 90° out of phase with the resultant loop voltage. The load of V_1 is a pair of untuned inductances L_6 and L_7, so that the output voltage of V_1 is 90° out of phase with the valve current. Hence the output voltage of V_1 is either in phase or 180° out of phase with the resultant loop voltage. Hence when coupled into the search coil circuit it either adds to or subtracts from the resultant loop voltage to give the cardioid pattern of reception. This method of sense coupling makes the vertical aerial more nearly resistive since a very large value of swamping resistor R can be used compared to that in example 1.

The capacitor C_4 at the input to V_1

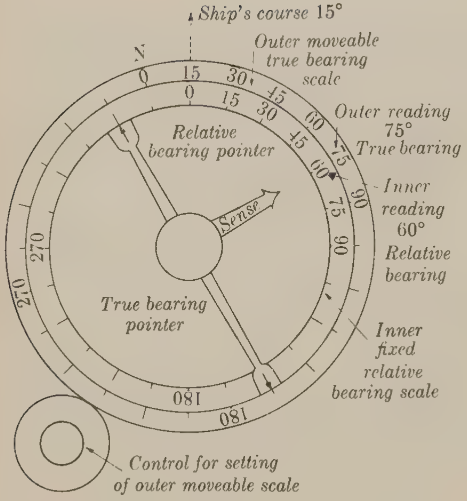

Ship's course 15°
N
Outer moveable true bearing scale
Outer reading 75° True bearing
Relative bearing pointer
Sense
Inner reading 60° Relative bearing
True bearing pointer
Inner fixed relative bearing scale
0
15
30
45
60
75
90
180
270
Control for setting of outer moveable scale

Fig. 14-37 True and relative bearing scales on goniometer.

is a 3-section differential capacitor used as a sense-voltage amplitude control. This capacitor provides two paths to ground for the aerial current, one through the grid-leak resistor R, and the other directly. The setting of C_4 thus controls the proportion of aerial current which flows in the grid-leak resistor R and hence the amplitude of the input voltage to V_1 which is across R. C_7 and R_1 provide cathode bias for V_1.

14.10 Position finding by d-f. The bearings normally required for navigation purposes are the relative bearings but in position finding true bearings are required. The goniometer has two scales, one outside the other. The inner scale of the goniometer gives the relative bearing, that is, the bearing of the transmitter relative to the ship's head. The bearing of the ship's course relative to the compass North-South direction on the outer scale is set opposite the zero of the inner scale. The outer scale gives the *true bearing*. In Fig. 14-37, for example, the bearing of the ship's course is 15° so the 15° mark on the outer scale is set opposite the 0° mark on the inner scale. Hence when the relative bearing of the transmitter is 60° the true bearing is read as 75° and is read on the outer scale. The true bearing scale is often electrically controlled by the ship's compass.

An example of the use of direction finders in position finding is shown in Fig. 14-38. The d-f operator on the ship X travelling north has taken bearings on three transmitters A, B and C which are land-based or ship transmitters, the positions of which are known. The true bearings of these transmitters as determined by the d-f operator are 270° for A, 330° for B and 45° for C. However it is necessary to find the bearing of the ship relative to each of the transmitters. In Fig. 14-39, if A has a bearing of 270° relative to X, then the bearing of X relative to A is 90°. Similarly the bearings

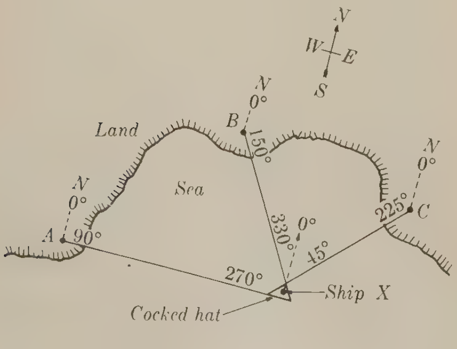


Fig. 14-38 Position finding by taking bearings on 3 transmitters of known positions.

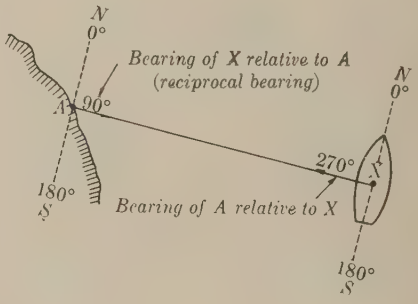

N 0°
Bearing of X relative to A (reciprocal bearing)
A
90°
180° S
Bearing of A relative to X
N 0°
270°
X
180° S

Fig. 14-39 Diagram to illustrate relationship between relative bearing and reciprocal bearing.

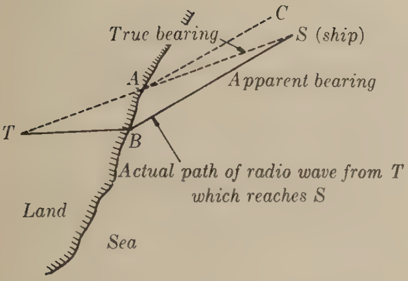


FIG.14-40 Error caused by shore effect.

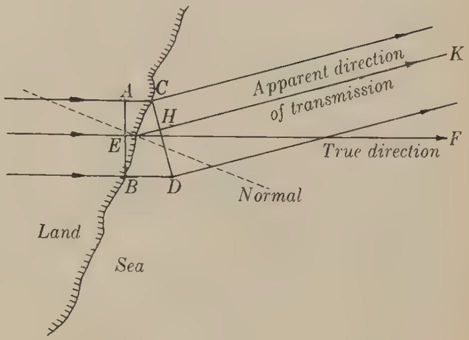


FIG.14-41 Diagram to illustrate reason for change in direction at shore line.

of the ship relative to B and C are 150° and 225° respectively. These new bearings, called *reciprocal bearings*, are plotted on the navigation chart as in Fig. 14-38. They should theoretically intersect at a point which represents the ship's position. In practice a small triangle called a *cocked hat* is obtained due to inaccuracy in the bearings obtained and the ship's position is taken to be at the centre of this triangle.

14.11 Polarization errors and other propagation effects. In sections 14.8 and 14.9 three types of d-f errors which can be corrected by proper design of equipment, and by calibration, are considered. In addition to these errors, there are other errors, caused by natural phenomena, for which no correction is possible. These errors are sometimes so large and unpredictable as to render direction finding useless. It is necessary therefore that a d-f operator recognize when these errors are present or are likely to occur so that bearings obtained at these times are not relied upon. The effects causing these errors are

(i) shore effect (refraction effect)
(ii) reflection effect
(iii) night effect
(iv) aeroplane effect.

A. Shore Effect

A cause of error in d-f is the change in direction of radio waves when passing across a shore line. This change in direction gives rise to an error in the observed bearing of the transmitter as illustrated in Fig. 14-40. Were no bending to take place the radio wave from the transmitter T would travel along the path TAS in reaching the ship S. Because of this refraction effect, however, this wave is bent towards C and does not reach S. The wave which does reach S follows another path TBS. Hence the observed or apparent bearing of T is that of the direction BS, while the true bearing is the direction TAS The error is represented in magnitude by the angle TSB.

This change in direction is analogous to the bending or refraction of light waves which occurs when light waves pass from one medium to a different medium. The refraction of light waves is explained by the existence of different velocities of light in different media. The refraction of radio waves is explained in a similar way by assuming a higher velocity over sea than over land.

In Fig. 14-41, AB is the front of a radio wave whose direction over land is EF. If the velocity is greater over water, then, in the time required for the wave front at A to reach C, the wave front at B has travelled a greater distance to D. The new wave front CD is not parallel to AB and has a new direction HK. The direction of bending is away from the normal to the boundary of the

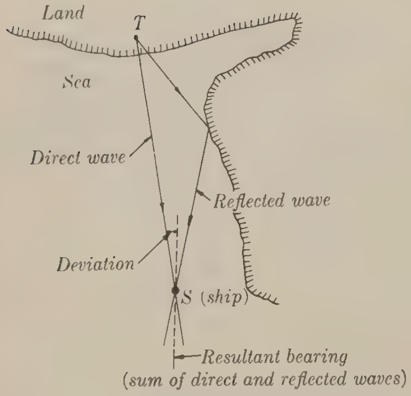


FIG. 14-42 Deviation caused by reflection effect.

land and water. If the wave is travelling from water to land the bending is toward the normal.

If the radio wave is travelling in a direction perpendicular to the coast, the wave fronts at A and B reach the coast at the same instant and no bending takes place. The amount of change in direction is greatest when the radio wave crosses the coast with a small angle between its path and the coast line.

This error caused by shore effect is never apparent to the operator and no correction can be applied. The error is avoided by locating land transmitters, which are to be used as d-f beacons by ships, as near the coast as possible. Then the path of the wave is almost entirely over water and little error occurs. Alternatively the operator should avoid taking bearings on land transmitters which are not near the coast.

B. Reflection Effect

Unreliable bearings are also obtained when a ship is in the neighbourhood of large land masses. In Fig. 14-42 the radio waves from a transmitter T are reflected towards S by the land masses nearby. The resultant of these reflected waves and the direct ray has a bearing which differs from the correct bearing. These errors can not be corrected for but the d-f operator must be aware of the possibility of their existence.

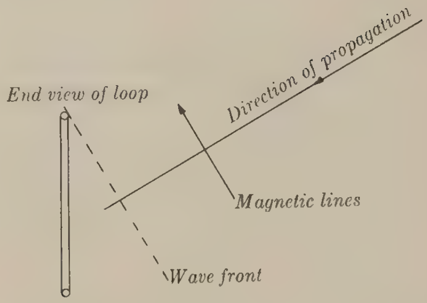
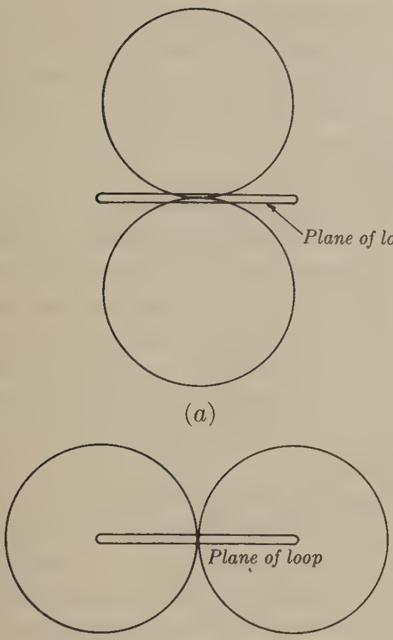


FIG. 14-43 Sky wave approaching a loop aerial with its plane vertical.

C. Night Effect

In considering the directional characteristics of a loop aerial, it is explained in earlier sections that when the loop is oriented with its plane perpendicular to the direction of propagation of a radio wave, the resultant loop e.m.f. is zero. This is true only if the wave is vertically polarized which is the case for a ground wave. Equal e.m.f.'s are induced in the vertical arms of the loop only acting in opposite directions around the loop and the loop e.m.f. is zero. The horizontal arms serve only to complete the loop circuit and to support the vertical arms. If a bearing is taken on a sky wave, however, the wave front is not parallel to the plane of the loop. In general, therefore, a sky wave is partly horizontally polarized. Thus e.m.f.'s are induced in the horizontal arms when the loop is oriented with its plane perpendicular to the direction of travel of the wave. Since the wave reaches the upper horizontal arm before the lower (Fig. 14-43) these two e.m.f.'s, although acting in opposite directions around the loop, do not cancel completely and there is a resultant loop e.m.f. However, with this orientation of the loop the resultant loop e.m.f. is expected to be zero. Hence an absolute minimum is not obtained. The orientation of the loop for which a minimum is obtained



(*a*)

(*b*)

FIG. 14-44 Horizontal polar diagram of loop aerial for sky wave which is (a) horizontally polarized (b) vertically polarized.

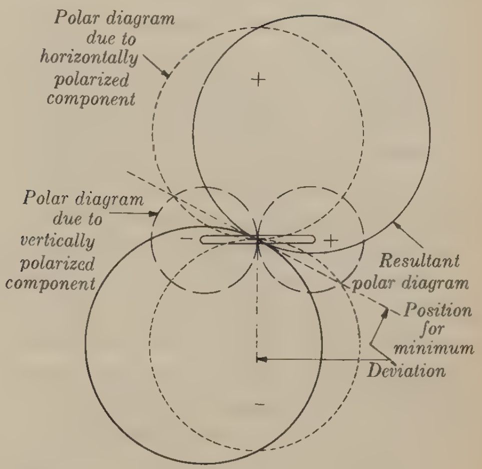

Polar diagram due to horizontally polarized component
+
Polar diagram due to vertically polarized component
−
+
Resultant polar diagram
Position for minimum
Deviation
−

FIG. 14-45 Polar diagram of loop aerial for sky wave with 30° angle of incidence, 50% horizontal polarization and no phase difference between horizontal and vertical electric field components.

depends on:

(i) the angle of incidence of the sky wave upon the loop
(ii) the percentage of horizontal polarization
(iii) the relative phases of the horizontal and vertical electric field components in the wave.

If the sky wave were entirely horizontally polarized, the horizontal polar diagram of reception for the loop would be as shown in Fig. 14-44 (a). When the plane of the loop is parallel to the line of propagation the horizontal electric lines are perpendicular to both the horizontal and vertical arms of the loop, and there is zero resultant loop e.m.f. since no e.m.f.'s are induced in any of the arms. When the plane of the loop is perpendicular to the line of propagation, e.m.f.'s are induced in the horizontal arms with a maximum phase difference so that the resultant loop e.m.f. is a maximum. The horizontal polar diagram of reception of a loop aerial for a completely vertically polarized wave is shown for comparison in Fig. 14-44 (b).

Fig. 14-45 shows the polar diagram of reception obtained for a sky wave having an angle of incidence of 30°, 50% horizontal polarization and no phase difference between the horizontal and vertical electric field components. It is seen that the figure-of-eight pattern obtained is rotated from the patterns which are obtained from vertically polarized or horizontally polarized waves only. The orientation of the loop for a minimum is thus widely different from the orientation for a ground wave of the same bearing. Hence a large error in bearing is introduced. Since a sky wave is obtained by reflection of a direct wave from the ionosphere, the horizontal and vertical electric field components are usually also out of phase. Thus the minima, besides being displaced, are indistinct. Since this effect occurs mainly at night it is called *night effect.* It is most noticeable at the medium r-f frequencies and in the hours just preceding and succeeding sunset and dawn. At these times the bearings obtained may be in error by as much as 30° or 40°. At other hours during the night the bearings obtained may have errors of only 3° to 5°. No correction can be applied to counteract the error in bearing caused by night effect. However, it is essential that the operator be aware of its presence, that is, know the symptoms of night effect which include the following.

(i) Indefinite minima 180° apart occur which are caused by an out-of-phase component of the wave.

(ii) Sharp minima not 180° apart occur caused by the presence of a horizontal electric field component in phase with the vertical electric field component. In some cases the minima may be 180° apart but displaced from the position of the correct bearing. In these cases the presence of night effect is not noticeable as far as the unknown transmission is concerned and the bearings obtained may be in error by as much as 90° depending on the angle of incidence of the reflected ray.

(iii) A combination of indefinite and displaced minima occurs. This is perhaps the most common symptom and may be identified by taking bearings on a transmitter of known bearing.

(iv) In position finding the cocked hat may be very large or the three bearings when plotted may not intersect at all on the chart.

(v) Fading of the signal strength occurs. If the receiving aerial is affected by both ground and sky wave transmission, the sky wave component varies due to changing refraction conditions at the ionized layer. Fading thus indicates that sky waves are being received and that night effect is possible.

(vi) The amount of balancing voltage needed to clear the minima is considerably larger than usual as indicated by a much larger amount of coupling than usual between the vertical aerial and the search coil circuit.

(vii) The position of the search coil for a minimum continually changes in an erratic manner. This also is due to changing refraction conditions.

In general any indication that sky waves are being received is an indication that night effect is present to some degree. Although no correction is possible an experienced operator learns to recognize the symptoms of sky wave reception and of night effect, and knows to what extent the bearings obtained are reliable. With normal sky wave reception at angles up to 15° with the horizontal, a loop aerial gives bearings accurate to within 5° or so.

D. Aeroplane Effect

Another effect which shows the same symptoms as night effect and causes similar errors is called *aeroplane effect.* This is observed when an aircraft transmission is being received. Because of the height of the aircraft and the fact that the direct radiation from its aerial reaches the ground as a sky wave, the wave contains a horizontal electric field component and night effect occurs. The bearing observed changes rapidly as the aircraft moves. Errors up to 40° result and the sign of the error is reversed when the aircraft is flying in the opposite direction. If the aircraft has a trailing aerial, the d-f apparatus locates the plane behind its actual position by as much as 20°. It can be shown that the d-f bearing obtained is the bearing of the point at which the line of the trailing aerial cuts the ground level.

E. Other Atmospheric Effects

Another atmospheric error is caused by the scattering of radiation by dust particles. This scattered radiation may reach the d-f apparatus from a widely different bearing than the direct radiation, and the resultant bearing will be in error. Fading of signal strength or erratic fluctuations in the zero position of the search coil are symptoms of this effect. The phase of the scattered radiation is in general different from that of the direct radiation when both are combined at the d-f system so that indistinct minima are also a symptom.

14.12 High-frequency direction finding (h-f d-f) and the Adcock aerial. At high frequencies of the order of a few Mc/s or more the range of the ground wave is small. Reception and direction finding over greater ranges is by means of sky waves. The non-vertical polarization of the sky waves causes such large errors in the bearings obtained using a closed loop that the closed loop aerial is useless in direction finding at these frequencies. These errors arise when the received wave contains a horizontal electric field component. Since these errors are unpredictable and can not be corrected for, a modified loop aerial, called the *Adcock aerial*, is used to avoid these errors. It has been explained that in the normal loop aerial it is the e.m.f.'s induced in the horizontal arms by a sky wave with a horizontal component of polarization which cause the error. The Adcock aerial is designed so that it is not affected by a horizontally polarized wave by reducing the effective length of the horizontal arms to a minimum or balancing out of the circuit any e.m.f.'s induced in them.

Fig. 14-46 shows the Adcock aerial in its simplest form, known as the *Adcock U aerial*. It consists of two vertical aerial arms connected at the bottom and inductively coupled to a receiver situated half-way between. Since the upper horizontal arm is omitted, the amount of e.m.f. induced by a horizontal electric field component is considerably reduced and the error due to night effect is less than that obtained with the closed loop aerial. The Adcock aerial may be used in a Bellini-Tosi system particularly where waves of fixed frequencies are received over long periods of time. Two Adcock aerials are fixed at right angles and connected to a radio goniometer as in the case of the complete loop aerials. However rotating Adcock aerials are simpler to set up and operate on land.

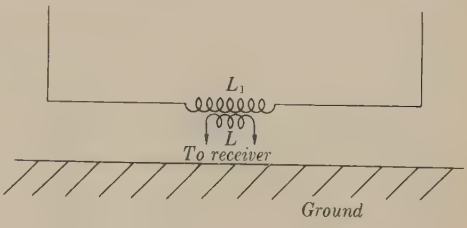


Fig. 14-46 Adcock U aerial.

Closed loop aerials usually consist of a number of turns to increase the amplitude of the loop e.m.f. This is not possible with the Adcock aerial. Hence in order to obtain a large enough loop e.m.f. to operate the receiver, the two vertical arms must be longer and more widely spaced than in the case of the closed loop aerial. The spacing may be as great as $\frac{\lambda}{4}$ at the upper end of the d-f station's frequency range. Because the Adcock aerial must have large dimensions its use is confined to land d-f stations, where its use is almost universal. The polar diagram of reception for the Adcock aerial is a figure-of-eight, the same as for the ordinary closed loop. With the Adcock aerial, as with the closed loop, only the e.m.f.'s induced in the vertical arms are combined to give the resultant loop e.m.f.'s. Hence there is no difference in circuit action.

The simple Adcock aerial shown in Fig. 14-46 eliminates the upper horizontal arm but still retains a lower horizontal arm made up of the two lead-in connec-

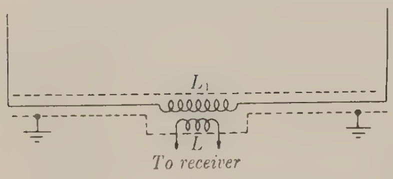


FIG.14-47 Shielded-U Adcock aerial.

tions from the vertical arms to the radio goniometer and receiver. The amount of pick-up by these connections is reduced by enclosing them in a metal shield that is connected to ground (Fig. 14-47). An aerial with this modification is called a *shielded-U Adcock aerial.* However, a disadvantage is that charges accumulate on the shield due to e.m.f.'s induced in it. The fields due to these charges couple the horizontal and vertical arms producing errors. To make the shielding more effective, particularly at higher frequencies, an artificial ground of higher conductivity is used consisting of a buried metal plate or wire network. A variation of the shielded-U is the *buried-U Adcock aerial* in which the horizontal section with its shield is buried to a depth of three feet or so.

In other forms of the Adcock aerial a balancing arrangement is used to cancel out the voltages induced in the horizontal arms. Fig. 14-48 shows the *elevated-H Adcock aerial.* The vertical arms are broken at the centres and connected together through the coupling coil L_1. Double lead-in wires have their connections to L_1 reversed. The currents i_1 and i_2 induced in the horizontal sections leading from one aerial are equal in amplitude but flow through L_1 in opposite directions and so cancel. However, the currents due to the e.m.f.'s induced in the vertical portions flow in the same direction through L_1 and so are coupled to the radio goniometer. The disadvantage of this type which gives a small error is that the radio goniometer and receiver must be located at some height above ground. If the receiver connections and the horizontal arms are located near ground level, a second loop aerial, consisting of the ground, the lower vertical sections and the horizontal section, is formed. The presence of this loop of different electrical characteristics unbalances the desired loop circuit.

Fig. 14-49 shows the *balanced Adcock aerial* which is similar to the *H*-type but has networks Z_1 and Z_2 replacing the lower vertical arms. The upper parts of the vertical arms are as long as the two separate sections were in the *H*-type so that the receiver can be located at ground level. The impedances of Z_1 and Z_2 are adjusted for an accurate balance and hence cancellation of the currents in the two horizontal lead-ins from each vertical arm occurs. Fig. 14-50 shows the *coupled Adcock aerial.* The degree of magnetic coupling in the transformers is chosen so the impedance of the horizontal arm circuit is high to all e.m.f.'s except those due to the vertical arms. The transformers are usually screened. Fig. 14-51 shows the *balanced-coupled Adcock aerial* which also permits

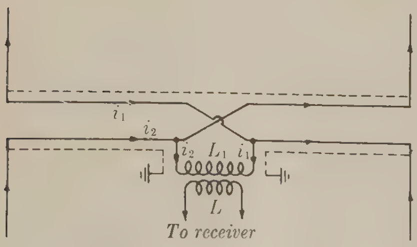


FIG.14-48 Elevated-H Adcock aerial.

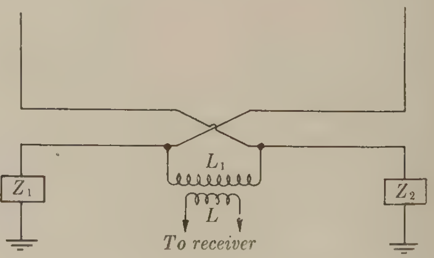


FIG.14-49 Balanced Adcock aerial.

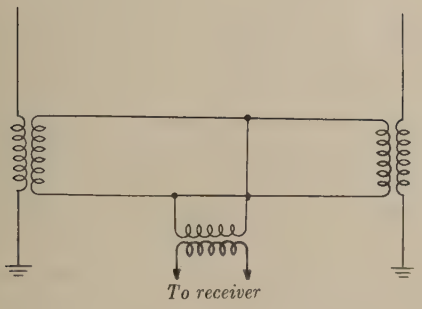


FIG. 14-50 Coupled Adcock aerial.

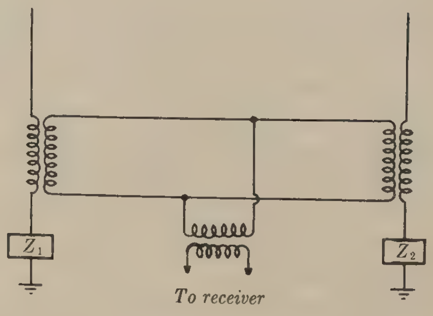


FIG. 14-51 Balanced-coupled Adcock aerial.

a ground level receiver location.

Fig. 14-52 shows the *transmission-line Adcock aerial.* The vertical aerial arms are coupled by means of matching transformers T_1 and T_2 to the buried transmission lines leading to the receiver. The primaries of T_1 and T_2 are shunted by networks which are chosen to obtain equality of output from each aerial.

The process of balancing an Adcock aerial may be accomplished by using a radiating oscillator located some distance away at such a height that its radiation reaches the aerial with a 45° angle of incidence. The lower portions of the two vertical arms which are adjustable in length are lengthened or shortened until the bearing of the oscillator obtained by the d-f apparatus equals the correct bearing. This balancing is usually carried out on a frequency at the middle of the frequency range of the d-f station.

To compare the performance of different types of d-f aerials a standard wave has been defined. This is a sky wave directed downward at an angle of 45° to the horizontal. The vertical and horizontal components of the electric field are thus equal in magnitude and this condition produces maximum error in bearing. The error obtained when taking a bearing on the standard wave using any given aerial is the *standard wave error* of the aerial. The standard wave error for some of the types of aerials already considered are:

(i)	complete loop aerial	35°
(ii)	Adcock U aerial	12°
(iii)	buried-U Adcock aerial	3°
(iv)	elevated-*H* Adcock aerial	4°
(v)	balanced Adcock aerial	6°
(vi)	coupled Adcock aerial	2°

Although the Adcock aerial is equivalent to only a single-turn loop, the pick-up of an Adcock aerial, as with a loop aerial, increases as the frequency increases. Hence since the Adcock aerial is used only at higher frequencies where sky waves occur, the pick-up is sufficient to operate a receiver effectively.

Adcock aerials are sometimes constructed so as to be capable of rotation in the vertical plane through 90°. Thus the vertical arms may be placed in any position from a horizontal position to a vertical position. This construction allows the Adcock aerial to follow or track the radiation from an aircraft. Since the electric field of a wave from

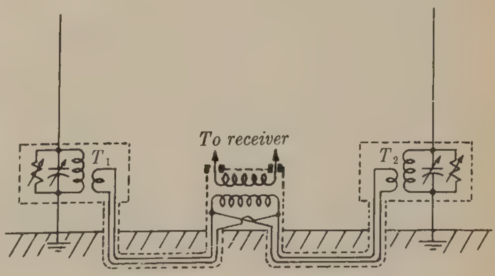


FIG. 14-52 Transmission-line Adcock aerial.

an aircraft is at an angle to the vertical, the vertical arms can be rotated so that they are parallel to the electric field. Under this condition maximum e.m.f. is induced in the vertical arms.

14.13 Comparison of Adcock (open loop) and closed loop aerials. A comparison of Adcock and closed loop aerials is given in the table at the bottom of this page.

14.14 The cathode-ray tube in direction finding. The cathode-ray tube is used in some h-f d-f equipment as a visual bearing indicator. Two methods of using it are considered in this section. In the first method the c.r.t. is used as an electrostatic goniometer replacing the radio goniometer. In the second the c.r.t. is used as an automatic bearing indicator replacing the earphones, loudspeaker or output meter and the goniometer scale.

A. The C.r.t. as a Goniometer

Fig. 14-53 illustrates in principle the way in which the c.r.t. is connected to replace the goniometer. The p-s loop of the fixed-frame aerial system is connected to the horizontal-deflection plates or X-plates of the c.r.t. and the f-a loop to the vertical-deflection plates or Y-plates. Thus any alternating voltage induced in the p-s loop causes a horizontal trace on the c.r.t. screen while any alternating voltage induced in the f-a loop causes a vertical trace. When alternating voltages are induced in both loops simultaneously the trace has an intermediate position on the screen, depending on the relative amplitudes of the voltages. These amplitudes in turn depend on the bearing of the signal being received.

If a signal arrives with a bearing of 0° or 180° an alternating voltage is induced

Comparison of Closed Loop and Adcock Aerials

Closed Loop Aerial	Adcock Aerial
1. Accurate only when receiving vertically polarized waves. Standard wave error 35°	Fairly accurate for waves with any polarization angle except the case in which the electric field is perpendicular to the dipoles
2. Accurate for waves at an angle of less than 5-10° relative to the horizontal	Accurate for waves at angles up to 60° relative to the horizontal
3. Accurate only for a range of 100 to 200 miles, which is the ground wave range	Range unlimited
4. Loop pick-up large due to the use of many turns	Adcock aerial pick-up equivalent to that of a single-turn loop
5. Useful only at low and medium r-f frequencies at which sky waves do not normally occur	May be used at high frequencies, (3-30 Mc/s), as well as at lower frequencies
6. Dimensions readily made small and so is used aboard ship	Large dimensions restrict its use to fixed shore stations
7. The errors not variable for any given frequency since its use is restricted to ground waves. Hence errors can be corrected for	Variable errors in general since it is used with sky waves and no complete correction can be applied

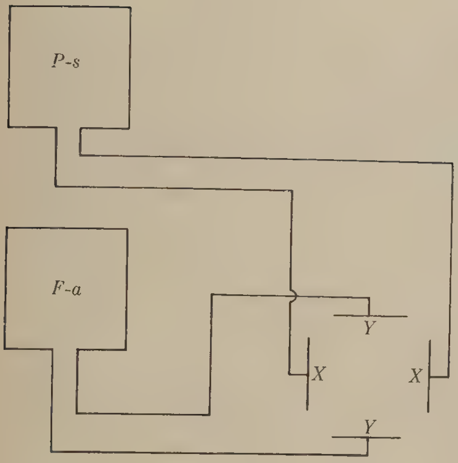
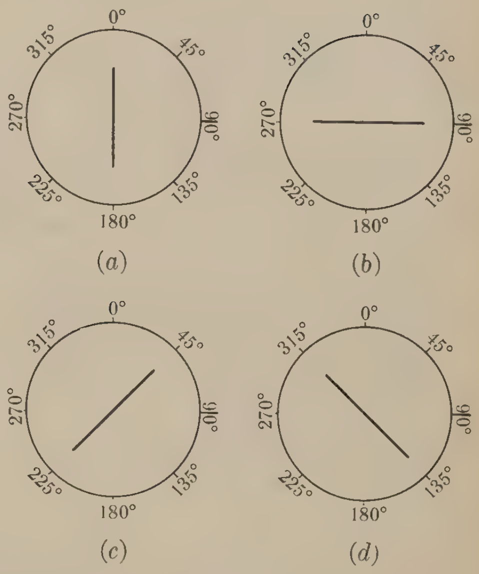


FIG. 14-53 Simple method of connecting c.r.t. as a goniometer.



(a) (b) (c) (d)

FIG. 14-54 Position of trace on c.r.t. used as a goniometer for signal with bearing of (a) 0° or 180° (b) 90° or 270° (c) 45° or 225° (d) 135° or 315°

in the f-a loop but not in the p-s loop. Hence the trace is a vertical straight line, Fig. 14-54 (a), its length depending on the field strength of the radio wave. Similarly a signal with a bearing of 90° or 270° causes a trace which is a horizontal straight line, Fig. 14-54 (b).

If the signal has an intermediate bearing the straight line trace is neither vertical nor horizontal but in between. Consider a signal with a bearing of 45°. The alternating voltages induced in the two loops are equal in amplitude and in phase. Hence the resultant trace is a straight line which makes an angle of 45° in a clockwise direction with the position of a vertical trace, Fig. 14-54 (c). If the signal bearing is 225° the voltages induced in the loops, although both reversed in phase, are still in phase with each other and equal in amplitude and the trace is again a straight line making an angle of 45° with the position of a vertical trace. Hence the straight line trace shown in Fig. 14-54 (c) is obtained for signal bearings of 45° or 225°. Thus a 180° ambiguity exists in the bearing pointed out by the trace on a scale arranged around the face of the tube. The scale reads from 0° to 360° as in the same diagram and so indicates the relative bearing of the signal.

If the bearing of the signal is 135° or 315°, the phase of the alternating voltage induced in the p-s loop is 180° different from its phase for signal bearings of 45° or 225°. Hence for either a 135° or 315° bearing the trace is a straight line making an angle of 45° in a counterclockwise direction from the position of a vertical trace, Fig. 14-54 (d).

To determine the correct bearing of the two bearings indicated, that is, the sense of the bearing, the voltage induced in a vertical aerial is connected to the control grid of the c.r.t. A simple method of connection is shown in Fig. 14-55. The vertical aerial must be resistive or at resonance so that the current through the coupling coil L is in phase with the voltage induced in the aerial and hence 90° out of phase with the loop voltages. The voltage applied to the control grid as a result of the coupling is 90° out of phase with the current in L and so is either in phase or

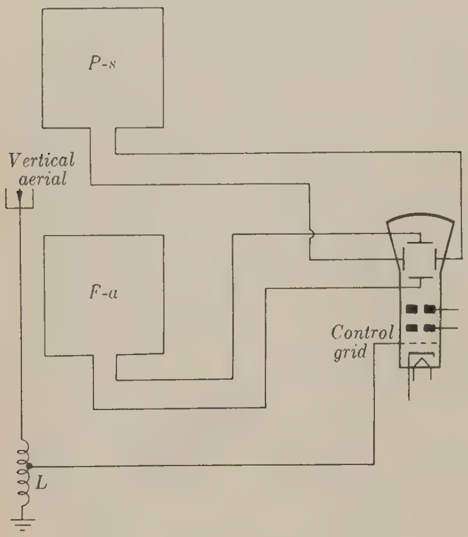


FIG. 14-55 Simple method of connecting vertical aerial to c.r.t. to give sense indication.

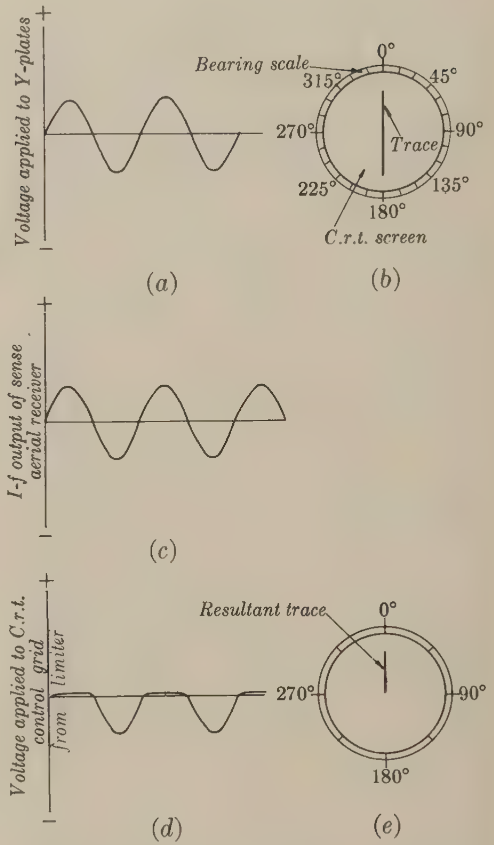

Voltage applied to Y-plates
+
−
(a)
Bearing scale
0°
45°
90°
135°
180°
225°
270°
315°
Trace
C.r.t. screen
(b)
I-f output of sense aerial receiver
+
−
(c)
Voltage applied to C.r.t. control grid from limiter
+
−
(d)
Resultant trace
0°
90°
180°
270°
(e)

FIG. 14-56 Diagrams to illustrate action of circuit of Fig. 14-55 for a signal with 0° bearing (a) Voltage applied to c.r.t. Y-plates (b) Trace due to this voltage (c) Output voltage of vertical aerial (d) Grid voltage (e) Resultant trace on c.r.t.

180° out of phase with the loop voltages. The grid bias is such that the a-c voltage applied to it, via a limiter, leaves the trace bright during the positive half cycle but blacks it out during the negative half cycle. Hence only one half of the trace is visible when the sense aerial is connected to the c.r.t. and only one bearing is thus indicated. The connections of the coil *L* to the circuit are made during calibration on signals of known bearing so that the visible half of the trace indicates the correct bearing or the sense of the signal. Fig. 14-56 illustrates the action for a signal with a bearing of 0°. The voltage applied to the c.r.t. *Y*-plates, Fig. 14-56 (a), produces a vertical deflection of the electron beam, Fig. 14-56 (b). The output voltage of the vertical aerial should be in phase, Fig. 14-56 (c). The resultant trace is shown in Fig. 14-56 (e).

The simple circuits of Figs. 14-53 and 14-55 are not practical because the amplitudes of the voltages induced in the loop and vertical aerials are too small to produce any noticeable deflection of the c.r.t. electron beam. In a practical circuit amplifiers must be inserted between the aerials and the c.r.t. A block diagram of such a circuit is shown in Fig. 14-57. Two superheterodyne receivers are used to amplify the aerial voltages, one for each loop aerial. Since the relative amplitudes and phases of the aerial voltages must not be changed before application to the c.r.t., each receiver must possess exactly equal gain and phase shift. The method by which this condition is achieved and maintained is discussed in a later paragraph under balancing procedure. The i-f outputs of the receivers are applied to the respective plates of the c.r.t. In addition the com-

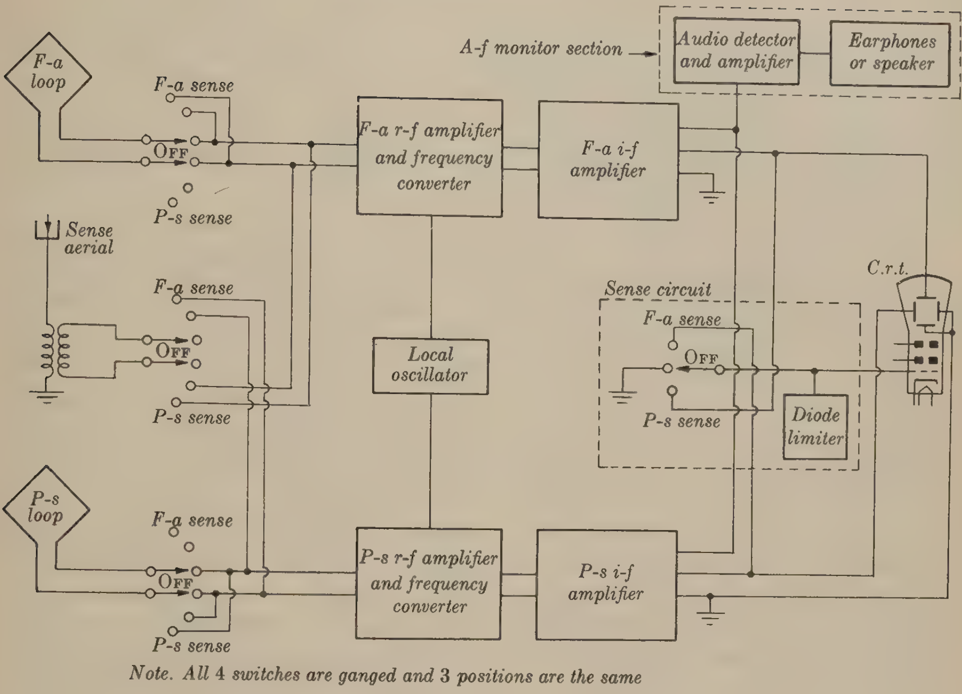


FIG. 14-57 Block diagram of practical circuit using c.r.t. as a goniometer.

bined i-f output of the receivers is connected to a second detector and the a-f output from this is connected to earphones or loudspeaker so that the signal message may be received or the identity of the transmitting station revealed. The section marked a-f monitor contains the second detector, a-f amplifier and phones or speaker. The sense aerial signal is amplified by first one receiver and then the other, the procedure being explained in a later paragraph.

The i-f output of the receiver amplifying the sense aerial signal is not applied directly to the control grid of the c.r.t. but is applied first to a rectifier circuit such as a diode limiter (Fig. 14-58). On the positive half cycle of sense aerial voltage the diode V conducts and the coupling capacitor C charges quickly through the low impedance of the diode. Hence only a small voltage appears across the grid-leak resistance R which is in parallel with V, and the c.r.t. grid voltage remains constant and slightly positive with respect to ground during the positive half cycle. The fixed grid bias is adjusted so that during this half cycle of sense aerial voltage the trace is brightly visible. On the negative half cycle of the sense aerial voltage the diode does not conduct and the capacitor C discharges through R causing a large voltage of the opposite polarity across R so that the control grid is sufficiently negative with respect to its cathode to make the trace invisible.

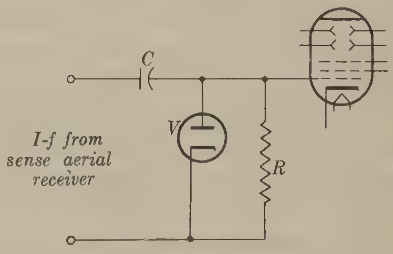
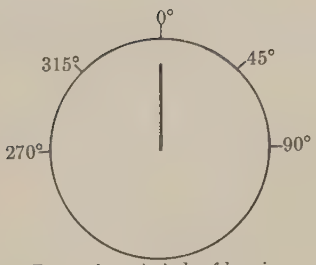
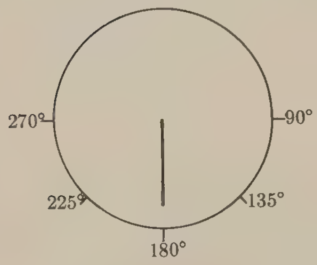


FIG. 14-58 Circuit of diode limiter used in block diagram of Fig. 14-57.



Forward semi-circle of bearings

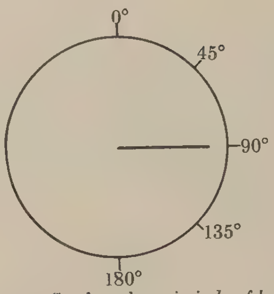
(*a*)



Aft semi-circle of bearings

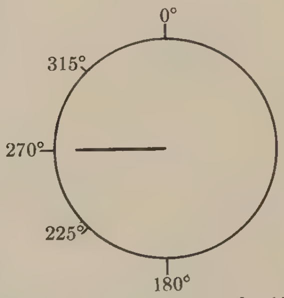
(*b*)

FIG.14-59 Sense trace on c.r.t. for (a) forward semi-circle of bearings (b) aft semi-circle of bearings.



Sense trace for Starboard semi-circle of bearings

(*a*)



Sense trace for Port semi-circle of bearings

(*b*)

FIG.14-60 Sense trace on c.r.t. for (a) starboard semi-circle of bearings (b) port semi-circle of bearings.

This action is also explained in terms of relative impedances. On the positive half cycle of sense aerial voltage the diode conducts and its impedance for i-f voltages is low compared to the impedance of C. Hence most of the input appears across C and very little across R, so that the control grid voltage does not change with respect to its cathode. On the negative half cycle the diode is non-conducting and its impedance is very high compared to the impedance of C. In addition the resistance of R is chosen so as to be large compared with the impedance of C. Hence most of the input negative half cycle of voltage appears across R and a large negative voltage pulse is thus applied to the control grid.

For sense determination there are two steps. In the first step the switch shown in Fig. 14-57 is turned to the f-a sense position. The p-s receiver is disconnected from the p-s loop and connected instead to the sense aerial to serve as a sense signal receiver. The f-a receiver remains connected to the f-a loop. There is no voltage applied to the X-plates of the c.r.t. and so the resultant trace is a straight line, one half of it being removed by the diode limiter, the action of which is explained above. For the forward semi-circle of bearings the sense and loop signals are arranged to be in phase so that the brilliance control grid is positive and the trace visible while it travels along the upper half of its sweep. When the beam travels along the lower half however the grid is negative enough to extinguish the trace. For the aft semi-circle the signals are in phase opposition and thus the trace is visible in the lower half of its path only. These conditions are illustrated in Fig. 14-59 (a) and (b).

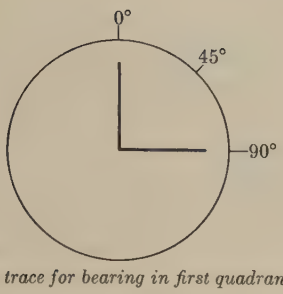


Fig.14-61 Sense traces for bearing in the first quadrant.

When the p-s sense position of the switch is used the f-a receiver is disconnected from the f-a loop and connected to the sense aerial. The p-s receiver remains connected to the p-s loop. Thus there is no voltage applied to the *Y*-plates of the c.r.t. and the trace is a horizontal straight line. For the starboard semi-circle of bearings the right hand half of the trace is visible while for the port semi-circle of bearings the left hand half of the trace is visible, Fig. 14-60 (a) and (b).

Usually both sense positions of the switch are used in rapid succession. The two half traces obtained enclose the quadrant in which the correct bearing lies (Fig. 14-61).

B. Procedure for Balancing the Loop and Sense Aerial Receivers

As previously pointed out the relative amplitudes and phases of the loop and sense aerial voltages must not be disturbed by the receivers. For example, if the signal bearing is 45° and the p-s receiver has a greater gain than the f-a receiver, the straight line trace produced on the screen will indicate a bearing greater than 45°. Also if one receiver shifts the phase of its input by a different amount than the other because of a difference in tuning, then the receiver outputs will be out of phase. In such a case the trace is not a straight line but an ellipse which is of little use in accurate bearing determination. Thus the total gain and total phase-shift (tuning) of the two loop receivers must be made exactly equal. These requirements are most easily satisfied if each stage of one receiver is matched for gain and phase-shift to the corresponding stage in the other. In addition the frequency-response characteristics of both receivers must be matched so that the match would be preserved for frequencies near the frequency tuned in.

To provide equal inputs to both receivers for the balancing procedure the output from a built-in test oscillator is connected to the input grids of both receivers by means of a switch. This oscillator automatically provides a continuous r-f signal at the frequency to which the receivers are tuned. Comparison of the stages is carried out by means of a number of metal links which when closed connect the signal grids of the corresponding stages. These links, called *comparator switches*, thus provide a common input to the two stages being matched. Hence if each corresponding pair of stages is compared starting from the output ends of the receivers the result will be a complete matching of the receivers. The phase-shifts of two stages are equalized by adjusting the tuning of their tuned circuits. Indication that this condition is reached is given by the production of a straight line trace of maximum length on the c.r.t. screen. The gains of two stages are equalized by adjusting the cathode bias of each. This balance condition is indicated by a 45° position of the straight line trace on the screen.

The exact balancing of the receivers is of little avail, however, if the two loop aerials and their circuits are unbalanced. In order to check their balance the test oscillator output is connected by another position of the same switch to the test aerial loop. This test loop is a small

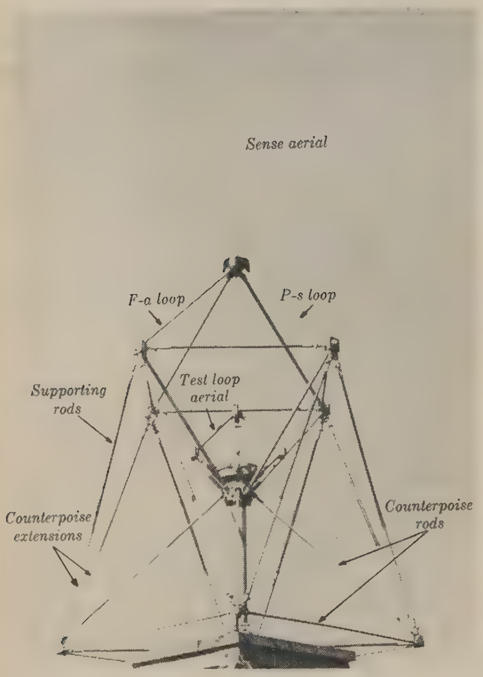


Fig.14-62 Loop aerial system with test loop.

screened loop placed inside the two loop aerials and at 45° to each of them (Fig. 14-62). This signal is fed at the same time to the sense aerial. Thus for test purposes this artificial signal imitates at the loop and sense aerial the electromagnetic field of an incoming radio wave having a 45° bearing. The phase-shift and output voltage magnitude of each loop aerial is adjusted until the trace on the screen is a straight line indicating a 45° bearing on the relative bearing scale around the face of the screen.

For each position of the sense switch adjustments are made in the sense circuit preceding the c.r.t. in order that the trace obtained is a straight line either vertical or horizontal and of minimum length. Since the i-f output which is applied to the sense circuit is constant in frequency regardless of what frequency is being received, the sense circuit adjustments can be pre-set and need only be checked once in a while.

C-r d-f is normally used at high frequencies for reasons given below under *C*. The aerial is mounted atop the mast with a counterpoise beneath to avoid re-radiation effects. A certain amount of error due to re-radiation is present and is corrected for from a calibration chart as in the case of the radio goniometer direction finder. C.r.t. radio goniometers are commonly used in conjunction with Adcock aerials in land installations as well as aboard ships.

C. *Advantages of Use of Cathode-Ray Tube in Direction Finders*

The advantage of a visual indicator over an aural indicator lies in the fact that a very brief signal, that is one of short duration, causes a trace on the c.r.t. screen of longer duration than the sound in a speaker due to the persistence of the screen. In the aural indicator such a brief signal would not be of any use since the search coil could not be moved quickly enough for a minimum. In addition when signal interference such as the reception of a signal of neighbouring frequency, atmospheric noise or fading is present, a visual reading is seen more readily than a minimum is heard in the aural type. Finally there is less chance for human error in the use of the c.r.t. With the aural type error arises in two places, first in finding a minimum and second in reading the scale. With the visual type only the latter error occurs.

D. The C.r.t. as an Automatic Bearing Indicator

In the second type of use of the c.r.t. in direction finding, the c.r.t. is used as an automatic bearing indicator device replacing the earphones or speaker of the aural type. A block diagram of a d-f system using an electromagnetic c.r.t. as an automatic bearing indicator is shown in Fig. 14-63. The two fixed

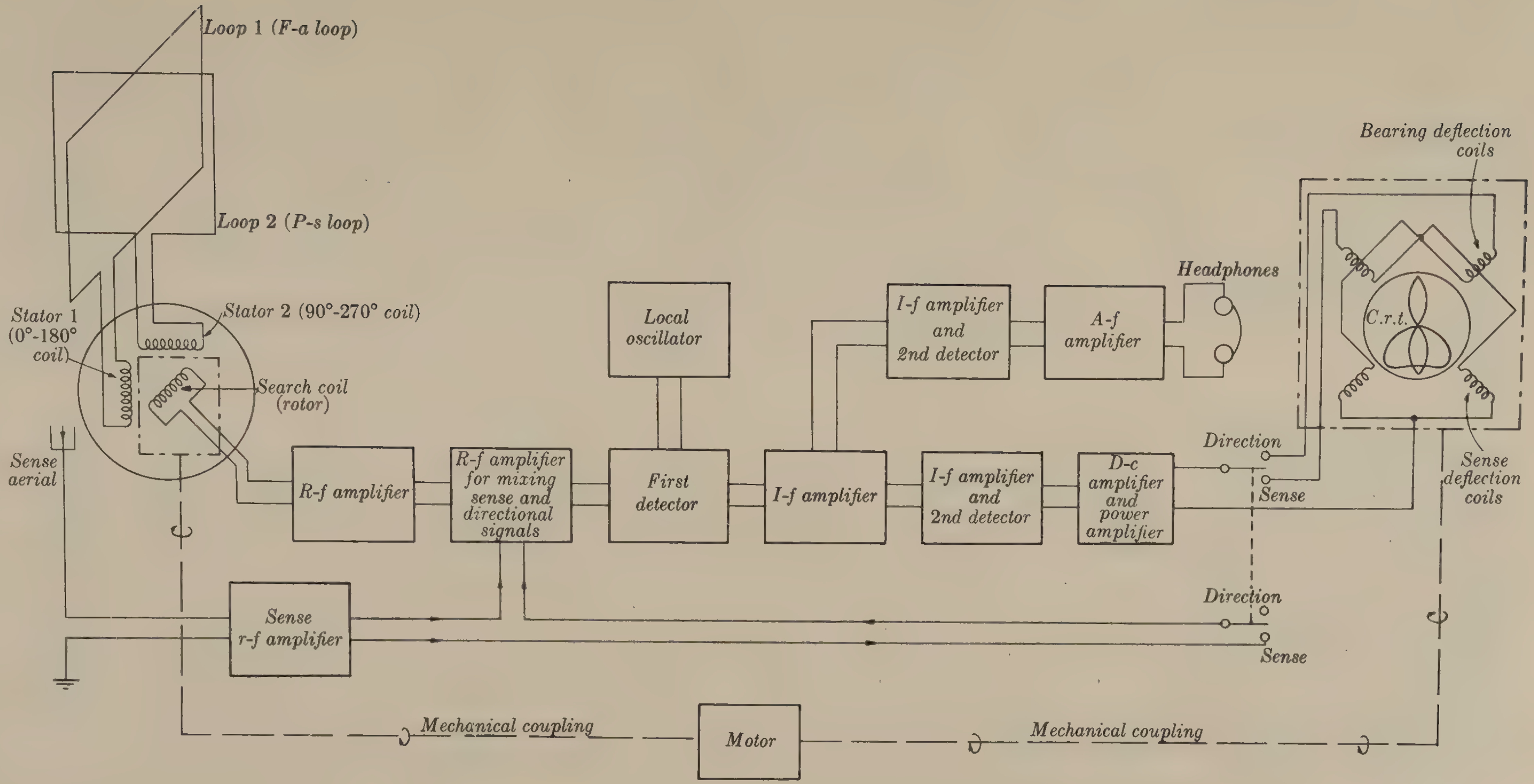


FIG.14-63 Block diagram of a d-f system using c.r t. as automatic bearing indicator.

aerial loops are connected to the respective fixed coils of the radio goniometer. The output of the search coil is applied to the input of a superheterodyne receiver as in any ordinary Bellini-Tosi system. The sense voltage is pre-amplified before being combined with the amplified loop voltage in a mixer stage. The output of this mixer stage is applied to the second detector to produce the i-f which is amplified. There are two second i-f amplifiers and two second detectors. The a-f output of one detector operates the earphones or speaker so that the message may be received or the transmitter identified. The i-f output of the other second detector is filtered of both i-f and a-f and amplified by a d-c amplifier which gives a negative d-c output voltage. This voltage, the amplitude of which depends on the amplitude of the input to the i-f stage, is applied to the control grid of a power amplifier. The anode current of this stage flows through the deflection coils of the c.r.t.

The equipment operates as an automatic response curve tracer, continually plotting the response of the search coil circuit to radio waves reaching the aerials. That is, it plots on the c.r.t. screen propeller-shaped and unilateral polar diagrams. The action is as follows.

When no signal is being received, the bias on the power amplifier is adjusted so that its anode current which flows through the deflection coils of the c.r.t. causes a field of such intensity that the spot on the screen is deflected to the outside edge of the screen. The search coil and deflection coils are rotated by the same motor at the same speed. Hence for zero signal the anode current of the power amplifier remains constant and the spot moves around the edge of the screen at the speed of the motor producing a glowing circular trace. If any signal is received from the search coil circuit, the negative d-c voltage applied to the power amplifier grid increases and its anode current decreases. Hence the magnetic field which controls the electron beam position is reduced and the glowing spot is deflected inward toward the centre of the screen by an amount which is proportional to the magnitude of the signal.

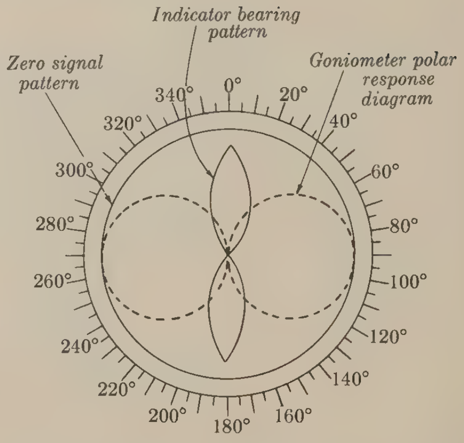


FIG. 14-64 Trace produced on c.r.t. automatic bearing indicator for signal with 0° bearing (sense aerial not connected).

In Fig. 14-64 the solid line illustrates the trace produced as the search coil is rotated through 360° for a signal with 0° bearing. The dotted line indicates the change in the output of the search coil circuit. When the search coil output is zero, the deflection coil current is a maximum and adjusted so that the glowing spot appears at the outside of the screen on the zero-signal trace. As the search coil is rotated its output voltage increases to a maximum. At the same time the deflection coil current is reduced to a minimum and the glowing spot moves progressively to the centre of the circle. The movement of the spot for the other 270° of rotation of the search coil (and the deflection coils) is shown. The resulting curve traced by the glowing spot is a propeller-shaped figure which is the reciprocal of the search coil response curve and points along the line of direction of the signal.

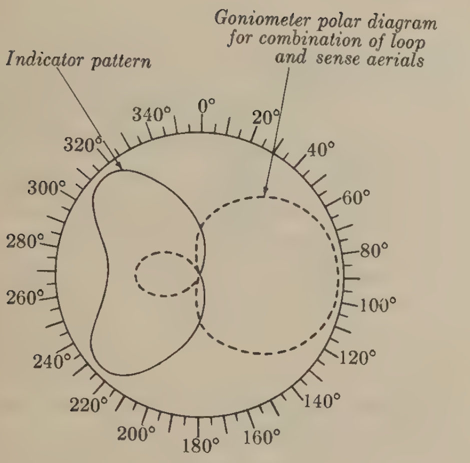


FIG. 14-65 Trace produced with sense aerial connected.

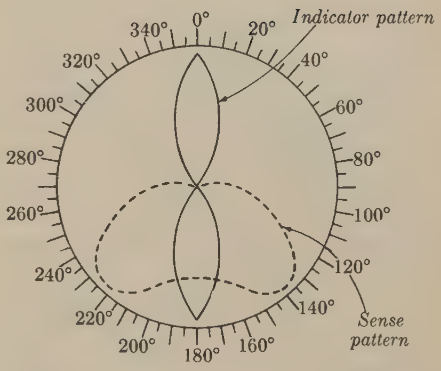


FIG. 14-66 Trace produced when sense pattern has desired orientation.

The sense pattern is obtained by combining the sense and search coil voltages. The dotted line in Fig. 14-65 shows the variation in response of the search coil and sense aerial circuits while the solid line indicates the pattern produced by the glowing spot if the p-a current were still connected to the bearing deflection coils. The output of the sense aerial is slightly smaller than that which would completely cancel the search coil voltage so that a small lobe is produced opposite the large lobe as shown by the dotted lines. Hence there are two minima present as the search coil is rotated through 360° which are not 180° apart. For the bearing of each minimum of the search coil output, the glowing spot appears at the outside of the screen. In any event a unilateral pattern is produced. This orientation of the pattern is not convenient for sense determination because this sense pattern indicates a bearing displaced from the true bearing by 90° while it ought to indicate one of the two bearings already indicated by the propeller pattern of Fig. 14-64. In practice this is overcome when switching to sense by arranging the circuit so that the p-a current is disconnected from the bearing deflection coils and connected to a pair of sense deflection coils which are placed around the tube 90° away from the other coils, but rotate at the same speed. The pattern then obtained is the same as that of Fig. 14-65 in shape but is rotated 90° away. Fig. 14-66 shows the sense pattern (dotted) compared to the d-f pattern (solid). The connections are arranged so that the sense pattern lies on the side of the screen opposite to the half of the d-f pattern which indicates the true bearing.

This type of c.r. d-f has the advantage over the one previously described in that only one receiver is necessary and hence no difficult balancing procedure is required.

14.15 Direction finding by directional transmission. The position of a ship or aircraft may be found using directional transmission. If a transmitter *T* of known position sends a known signal in only one known direction (Fig. 14-67) then the reception of this signal indicates that the receiver *R* lies along the direction of transmission. This principle is the basis of the following systems of direction finding by means of directional transmission.

(i) Revolving radio beacons
(ii) The radio range
(iii) Blind-landing devices.

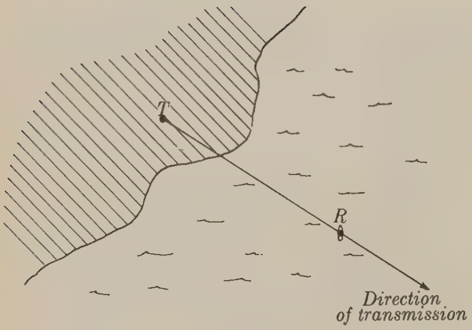


Fig.14-67 Direction finding using directional transmission.

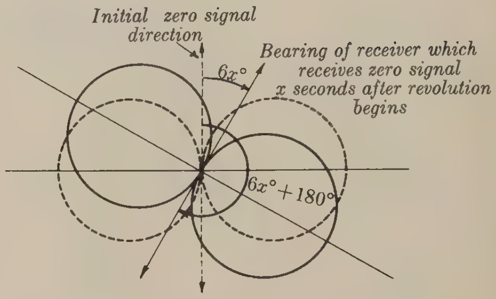

Initial zero signal direction
6x°
Bearing of receiver which receives zero signal x seconds after revolution begins
6x°+180°

Fig.14-69 Position of radiation pattern when receiver receives zero signal.

A. *Revolving Radio Beacons*

In one system of this type the transmitter has a radiation pattern consisting of a figure-of-eight which is caused to rotate continuously through 360° about the position of the transmitter, with a pause of known duration before the next revolution. Suppose the rate of rotation is once per minute. At any instant there is zero radiation along two directions 180° apart, Fig. 14-68 (a). Fifteen seconds later the directions along which there is zero radiation are 90° away, Fig. 14-68 (b), and so on. The system is set up so that the beam begins a 360° rotation with zero radiation along two known directions, north and south for example, as in Fig. 14-69. A given signal such as the Morse equivalent of the letter N is sent during the revolution. When the operator of a receiver hears this characteristic signal begin, he starts a stop watch. After a time x seconds depending on his location he hears zero signal. Since the rate of rotation of the beam is 6° per second the bearing of the receiver must be $6x°$ or $6x°+180°$ Enough is usually known of the probable direction of the transmitter to remove the 180° ambiguity. If the receiver is along the initial direction of zero signal the operator could not hear the signal at the beginning of the revolution. To avoid this difficulty a second signal, say the Morse equivalent of the letter A, is sent out when the direction for zero signal is east and west.

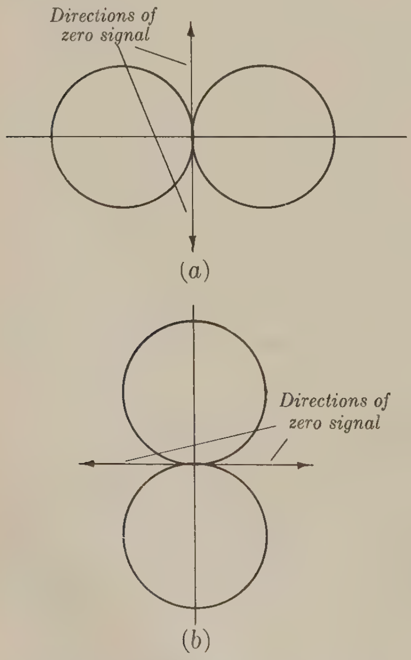


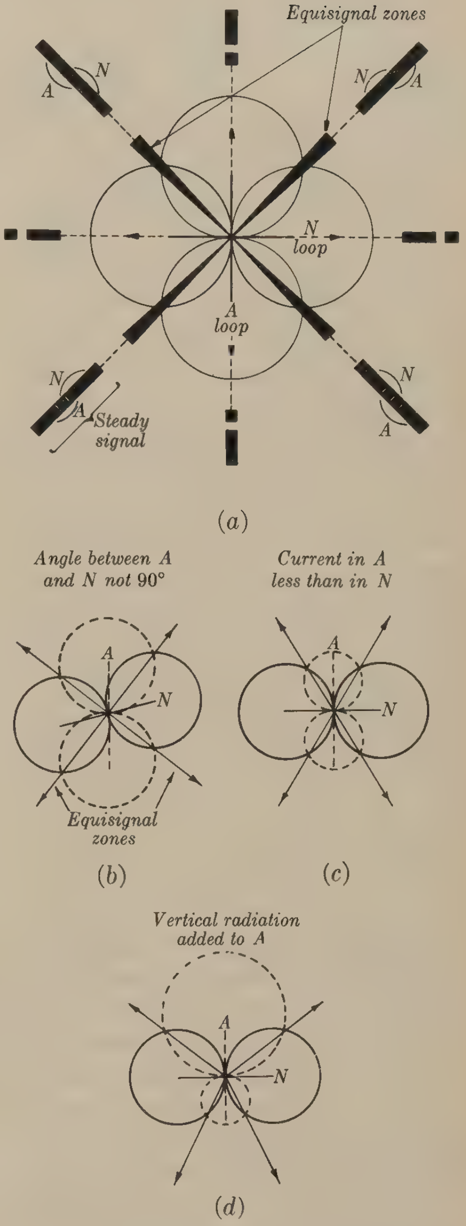
Fig.14-68 Radiation pattern of revolving radio beacon (a) at one instant (b) 15 seconds later.

B. *The Radio Range*

The radio range is used to indicate a predetermined path which aircraft may follow at night or under conditions of poor visibility. The on course condition is indicated by means of aural signals or visual instrument indications. A radio range transmitter aerial usually consists of two crossed loops A and B, Fig. 14-78 (a). Each loop is separately energized to radiate a characteristic signal. For example, one sends the Morse equivalent of the letter A (· -) and the other that for N (- ·). The directional patterns are also crossed, Fig. 14-70 (a). The aerials are phased so that the dit of the letter A occurs between the dah and dit of the letter N, that is, the signals chosen are interlocking. Along any one of the four directions shaded in on the diagram, an aircraft receives the two signals with equal intensity so that a continuous dah of sound is heard in speaker or headphones. If the aircraft is to either side of these directions, either the signal A or the signal N predominates. The aircraft is altered in course until the continuous dah is heard. Now the aircraft can follow the course, shaded in the diagram, to the landing field situated at the centre.

By varying the angle between the loops or the relative amplitudes of the loop currents, the directions of the courses shaded in can be varied, Fig. 14-70 (b), (c) and (d). These directions are called *equisignal directions or zones,* and each usually has a width of a few degrees. A goniometer arrangement may be used to shift these zones without changing the physical orientation of the loops.

Another form of radio range gives a visual indication on the instrument panel of the on course condition. In this system the two loops each radiate carrier waves of the same frequencies but modulated at differing audio frequencies. For example, the modulation frequency of one may be 65 c/s and of the other 85 c/s. If the aerial currents have the same amplitude and the degree of modulation is the same equisignal zones are produced. If the aircraft is flying in one of these zones, the detected voltages resulting from the reception of the two modulated carrier waves have the same amplitudes. Each detected low-frequency voltage is applied to a separate metal


Equisignal zones
N
A
N
A
N loop
A loop
N
A
Steady signal
N
A
(a)
Angle between A and N not 90°
A
N
Equisignal zones
(b)
Current in A less than in N
A
N
(c)
Vertical radiation added to A
A
N
(d)

Note. Dotted curves refer to A loop
Solid curves refer to N loop

FIG. 14-70 Equisignal zones in a radio range.

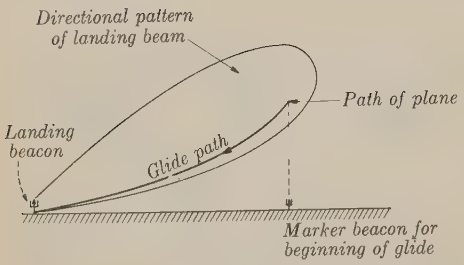

Directional pattern of landing beam
Path of plane
Landing beacon
Glide path
Marker beacon for beginning of glide

Fig. 14-71 Directional pattern for a blind-landing radio device.

reed situated behind one of the dials of the instrument panel. Each reed vibrates mechanically at the respective modulating frequency, one being constructed to have a mechanical resonant vibration frequency of 65 c/s and the other 85 c/s. When the voltage amplitudes are equal the amplitudes of vibration are also equal. If the tips of the reeds are painted white and the dial black, two white lines of equal length are observed. If the aircraft goes off course, one line becomes longer since one voltage will have greater amplitude than the other. The aircraft can therefore make the proper change in direction to get back on course.

C. *Blind-Landing Devices*

There are a number of radio devices to enable an aircraft to make a landing under conditions of poor visibility such as in a fog or at night. One device uses two or more portable transmitters stationed at varying distances from the field along the desired direction of landing. The aircraft obtains by wireless from the control tower of the landing field the direction in which to approach the field so as to pass over these transmitters, called *marker beacons*, and also the distance of each beacon from the field. As the aircraft approaches the characteristic signal of each beacon is received in turn. The aircraft then makes the correct changes in altitude and speed in order to land safely.

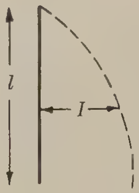


Fig. 14-72 Distribution of current along a $\frac{\lambda}{4}$ vertical aerial.

In another system an u-h-f radio beacon is used, with its direction of transmission aimed upwards at a small angle with the ground. The aircraft then flies down this beam by changing altitude so as to obtain constant intensity of the output of its receiver. The intensity and angle of radiation are designed so that if the aircraft maintains this condition it follows the path shown in Fig. 14-71.

14.16 Applied mathematics.

A. *Effective Height of a Simple Vertical $\frac{\lambda}{4}$ Aerial*

Fig. 14-72 shows the distribution of the standing wave of current along a simple vertical $\frac{\lambda}{4}$ aerial of length l. The radiation from this aerial varies from point to point along the aerial because of the unequal distribution of current. The radiation is greatest at the lower end where the current amplitude is greatest and equal to I, and least at the upper end. It can be shown that the average value of the current amplitude is $\frac{2I}{\pi}$. Hence an aerial with a length of $\frac{2}{\pi}\times l$ and a uniform current of amplitude I gives the same total radiation as the aerial shown of length l.

The expression $\frac{2l}{\pi}$ is called the *effective height* of the aerial. It represents the length of the vertical aerial which produces the same field intensity at some given position away from the aerial as is produced by the aerial shown in Fig. 14-72, but carrying a uniform current of amplitude equal to the maximum amplitude of current in the aerial of Fig. 14-72. Since π is approximately 3, the effective height of a simple quarter-wave vertical aerial is about $\frac{2}{3}$ of its actual length.

This relationship is also true for receiving aerials. The e.m.f. induced in an aerial is the product of its effective height and the field strength of the wave. Thus if the field strength at a quarter-wave vertical aerial of height h meters is x volts per meter the e.m.f. induced in the aerial is

$$x \times \frac{2h}{\pi} = \frac{2xh}{\pi} \text{ volts.}$$

The effective height of a simple vertical aerial, it is noted from this relation, is independent of the frequency of the radio wave.

B. *Effective Height of a Simple Loop Aerial*

It can be shown that the amplitude of the resultant e.m.f. set up in a rectangular loop, the plane of which is vertical, by an incident vertically polarized radio wave is given by the following equation.

$$E_L = 2EhN \sin\left(\frac{\pi d \cos\theta}{\lambda}\right) \qquad (1)$$

where E_L = amplitude of the resultant loop e.m.f. in volts

E = field strength of radio wave in volts per metre

h = height of loop, that is the length of each vertical arm, in metres

d = width of loop, that is the distance between the vertical arms in metres

N = number of turns in the loop,

λ = wavelength of radio wave in metres

and θ = angle between the plane of the loop and the direction of travel of the radio wave.

In practical loops the width is small compared with a half wavelength so that $\frac{\sin \pi d \cos\theta}{\lambda}$ may be written as $\frac{\pi d \cos\theta}{\lambda}$ without appreciable error. Hence the equation (1) becomes,

$$E_L = \frac{2EhN\pi D \cos\theta}{\lambda}$$

$$= \frac{2\pi N\, hd \cos\theta}{\lambda} \times E$$

But hd is the area A of the loop in sq. metres so that the equation may be written as,

$$E_L = \frac{2\pi NA \cos\theta}{\lambda} \times E \qquad (2)$$

The expression for E_L has a maximum value when $\cos\theta = 1$, that is, when $\theta = 0°$ or the plane of the loop is parallel to the direction of travel of the radio wave. The maximum value of E_L, denoted by $E_{L\ max}$, is then given by

$$E_{L\ max} = \frac{2\pi NA}{\lambda} \times E \qquad (3)$$

Since the expression $\frac{2\pi NA}{\lambda}$ when multiplied by E, the field strength of the wave, gives the maximum amplitude of the resultant loop e.m.f., it is called the *maximum effective height* or *effective height* of the loop aerial and is expressed in metres. Because $\lambda = \frac{v}{f}$ where v is the velocity of the wave in metres per second and f is the frequency in cycles per second, the effective height of a loop aerial is equal to $\frac{2\pi fAN}{v}$ in metres. For a single turn loop $N = 1$ and the effective height in metres is then given by,

$$h_{eff} = \frac{2\pi fA}{v} \qquad (4)$$

It is noted that the effective height of a loop aerial depends directly on the frequency of the radio wave. Thus the pick-up of a loop aerial increases with frequency. This is because the width d of the loop is increasing towards becoming a half wavelength.

C. Comparison of the Effective Heights of a Loop Aerial and a Vertical Quarter-Wave Aerial

If the effective height of the single-turn loop aerial is denoted by h_L and that of the vertical quarter-wave aerial (of height h) by h_V then,

$$\frac{\text{Effective height of loop aerial}}{\text{Effective height of vertical aerial}} = \frac{h_L}{h_V}$$

Substituting for h_L and h_V,

$$\begin{aligned}\frac{h_L}{h_V} &= \frac{2\pi f dh/v}{2h/\pi}\\ &= \frac{2\pi f dh}{v}\times\frac{\pi}{2h}\\ &= \frac{\pi^2 f d}{v}\end{aligned}$$

At a frequency of 500 Kc/s, and with a loop of width 1 metre, this ratio becomes,

$$\begin{aligned}\frac{h_L}{h_V} &= \frac{\pi^2\times 500\times 10^3\times 1}{3\times 10^8}\\ &= .016\\ &\doteqdot \frac{1}{60}\end{aligned}$$

Thus a loop aerial of one turn has only $\frac{1}{60}$ as much e.m.f. induced in it as a vertical quarter-wave aerial of the same physical height. The loop aerial e.m.f. is increased by using more than one turn. For example a 6-turn loop is $\frac{6}{60}$ or $\frac{1}{10}$ as effective as the vertical aerial. Thus it is necessary when loop aerials are used that the receiver be very sensitive, particularly at lower frequencies.

D. Effective Height of the Adcock Aerial

The effective height of the Adcock aerial is given by

$$h_{eff} = \frac{2\pi l s \cos\theta}{\lambda}$$

where l = length of vertical arms in metres

s = spacing of vertical arms in metres

λ = wavelength in metres

θ = angle between plane of aerial and the direction of propagation of the wave in degrees.

If E is the amplitude of the wave field strength at the aerial in volts, then the resultant output voltage of the loop in volts is equal to $E\times$effective height.

This quantity is a maximum when $\cos\theta = 1$ or $\theta = 0°$, that is when the plane of the aerial is along the wave direction. Thus the maximum amplitude of the aerial's output voltage is

$$E\times\frac{2\pi A}{\lambda}\text{ volts}$$

where $A = l\times s$, that is, the area of the loop in sq. metres.

14.17 Questions and problems.

1. Explain, using diagrams, how the simple loop aerial is used in direction finding.
2. Explain why the loop e.m.f. is a maximum when the width of the loop is equal to a half wavelength.
3. Explain the 90° phase difference between the resultant loop e.m.f. and the incident radio wave.
4. Why is there a 180° ambiguity in the apparent direction of the transmitter when using the loop aerial?
5. Why is the position of the loop for zero or minimum loop e.m.f. used rather than the position for maximum loop e.m.f.? Draw a diagram to aid the explanation.

6. How is the use of the loop affected if the incident radio wave is not vertically polarized?
7. When is it necessary to determine the sense of a bearing and when is it not necessary to do so?
8. Why must there be a 90° phase shift when coupling the sense aerial voltage into the loop voltage circuit?
9. Explain the changes which take place in the combined sense and loop aerial voltage as the loop aerial is rotated through 360°. Draw and describe the shape of the resultant polar diagram of reception.
10. Show how the polar diagrams of reception for a vertical aerial and a loop aerial are combined to give the polar diagram for the combination.
11. Why is the combination of sense and loop aerials not used to read a bearing accurately? What does the combination determine?
12. Explain the fact that a zero obtained using the combination finds the loop aerial in a position 90° different from that when a zero is obtained with the loop alone.
13. What are the disadvantages of a rotating loop aerial?
14. Explain the arrangement and components of a Bellini-Tosi aerial system.
15. What are the advantages of the Bellini-Tosi system?
16. Explain the action of a Bellini-Tosi aerial system if a signal is received from a bearing of 0°.
17. Explain the action for signals having the following bearings: 90°, 180°, 270°, 45°, 135°, 50°.
18. List the errors which arise in direction finding.
19. What causes the error known as vertical effect?
20. Describe aerial circuit arrangements which help to eliminate vertical effect.
21. Why is the metal shield often used about the aerial divided and joined by an insulator?
22. Explain how errors are caused by re-radiation.
23. Describe how the parts of a ship act as re-radiators. Describe the two kinds of re-radiators which exist.
24. Explain why the error caused by re-radiation from vertical radiators is called semi-circular error.
25. Why can errors due to re-radiation not be eliminated?
26. Explain how semi-circular error is corrected for.
27. Why does re-radiation blur the zeros?
28. Explain how the poor definition of the zeros is eliminated.
29. How can the one vertical aerial be used both for sense determination and for producing definite zeros?
30. Explain clearly the phase relationships which must exist
    (a) when the vertical aerial is used for sense determination
    (b) when it is used as a source of balancing voltage.
31. Explain why the error caused by re-radiation from loop radiators is called quadrantal error.
32. How is quadrantal error reduced by
    (a) changes in size of the loop aerial
    (b) use of a calibration choke?
33. Explain the construction and operation of a compensating loop.
34. Why is it not possible to have two correction curves, one for semi-circular error and one for quadrantal error?

35. What is the shape of the single correction chart?

36. Under what circumstances would the single correction curve look like the theoretical correction curve for either semi-circular or quadrantal error?

37. Explain the arrangement and action of the aerial circuit shown in Fig. 14-32, mentioning each circuit component.

38. Give an explanation corresponding to that of question 37 for the circuit shown in Fig. 14-34.

39. Describe the advantages in the use of the circuit of Fig. 14-34 over that of Fig. 14-32.

40. List the steps required in the direction finding procedure using the aerial circuit of Fig. 14-34.

41. Explain how the position of a ship or aircraft is determined by radio providing the locations of two or more transmitters which can be received are known.

42. Explain how the position of an unlocated transmitter is determined by radio by two or more ships whose positions are known.

43. Explain how the goniometer scales are arranged to give either relative or true bearing readings.

44. What is meant by the term cocked hat in d-f?

45. Name and describe briefly three errors which are called atmospheric errors.

46. (a) Describe how each of the errors listed under question 45 are recognized.
    (b) What action should be taken by the d-f operator when these errors are present?

47. Explain how night effect causes errors in d-f.

48. (a) Describe the symptoms of night effect.
    (b) When does this effect cause the greatest errors?

49. Why is the simple loop aerial not useful for high-frequency d-f?

50. Explain how an Adcock aerial can be used for high-frequency d-f.

51. Why must an Adcock aerial have larger dimensions than a simple loop aerial?

52. How does a shielded-U Adcock aerial differ from a simple Adcock U aerial in construction and advantages?

53. Explain the construction, advantage and disadvantage of the elevated-H Adcock aerial.

54. How is the disadvantage of the elevated-H Adcock aerial overcome in the balanced Adcock aerial?

55. Explain how a balanced-coupled Adcock aerial is balanced.

56. What is meant by the term standard wave error as applied to aerials?

57. Compare the standard wave errors of a simple loop aerial and some of the Adcock aerials.

58. Make a table to compare the Adcock aerial and the simple loop aerial with respect to
    (a) accuracy for ground waves and sky waves
    (b) range
    (c) pick-up or sensitivity
    (d) dimensions
    (e) usefulness.

59. Explain how a c.r.t. is used as a goniometer.

60. Explain what kind of traces are produced by signals of various bearings.

61. Describe how sense is determined when the c.r.t. is used as a goniometer.
62. (a) Explain the difficulties which arise in amplifying the aerial voltages before they are applied to the c.r.t.
    (b) How are they overcome?
63. Draw and label a block diagram of a practical direction finder using a c.r.t. as the goniometer.
64. Explain the action of the diode limiter in Fig. 14-58.
65. Describe the receiver balancing procedure for the direction finder described in the answer to question 63.
66. Describe the advantages of the use of a c.r.t. as a goniometer.
67. Explain how a c.r.t. is used as an automatic bearing indicating device in conjunction with a Bellini-Tosi aerial system.
68. Draw a block diagram of a direction finder which uses the c.r.t. as an automatic bearing indicator.
69. Describe the trace on the c.r.t. obtained when the search coil is rotated through 360°
    (a) without the sense aerial connected
    (b) with the sense aerial connected.
70. What is the advantage in using the c.r.t. as an automatic bearing indicator rather than as a goniometer?
71. Explain how direction finding can be carried out by directional transmission.
72. Describe the direction finding procedure with a revolving radio beacon.
73. Explain the way in which a radio range is used to guide aircraft.
74. Describe the operation of a radio blind-landing system.
75. Derive the formula for the effective height of a loop aerial having N turns, explaining the units normally used.
76. Use the formula of question 75 to find the voltage induced in a loop aerial given the following information:

    width of loop aerial = 1.25 metres
    height of loop aerial = 1.25 metres
    field strength of wave = 10 volts per metre
    number of turns in loop = 8
    frequency of wave = 300 Kc/s.

    Find the loop e.m.f. when its plane
    (a) is parallel to the plane of wave direction
    (b) is perpendicular to the plane of wave direction
    (c) makes an angle of 5° with the plane of wave direction
    (d) makes an angle of 10° with the plane of wave direction
    (e) makes an angle of 85° with the plane of wave direction
    (f) makes an angle of 80° with the plane of wave direction.
77. What conclusions are derived from a comparison of the results of question 76?
78. The dimensions of a loop aerial are to be 75 cm. by 75 cm. How many turns are required to give a loop e.m.f. of 1.4 microvolts maximum amplitude if the field strength of the wave is 30 microvolts per metre and its frequency is 1 Mc/s?
79. Repeat question 76 (a) for an Adcock aerial of the same dimensions. Compare the results with that obtained from question 76 (a).
80. Find the maximum amplitude of the e.m.f. induced in a vertical quarter-wave aerial under the conditions of question 76, and having a length equal to the height of the loop aerial of question 76.
81. Compare the results obtained from questions 76, 79 and 80.

CHAPTER XV

PULSE CIRCUITS

The term *pulse circuits* refers here to a wide range of circuits used in radar and television equipments. These circuits differ widely from conventional oscillators, amplifiers, rectifiers and detectors used in radio communications equipment in which the waveform of the operating voltage is usually sinusoidal or a simple combination of sinusoidal waves. Some circuits are used to develop square, sawtooth, trapezoidal and peaked waves of voltage which are required in timing, indicating and modulating circuits of radar equipment. The operating conditions of the valves in most cases range from cut-off to saturation during each cycle and do not fall into the simple classification of Class A, B and C operation. The circuits therefore are named for the function they perform rather than their type of operation. A circuit which limits the value to which a voltage may rise or fall is called a *limiter*. Two main classes of circuits which are included in this definition are treated under the sub-headings of clipping circuits and clamping circuits.

15.1 Clipping Circuits. A *clipping circuit* is one which limits the amplitude of a voltage by the removal of part of the waveform. Either the positive or negative extreme, or even both, may be limited. Several types of clipping circuits are described in the following paragraphs. The input to each circuit has a sinusoidal waveform but the operation is similar for any shape of a-c input waveform.

A. Series-Diode Clipping

The circuit of Fig. 15-1 is used to clip off the negative half of the input waveform. The voltage to be limited is applied to a diode valve in series with a resistor R. The output voltage is developed across the resistor. During the positive half cycle of applied voltage the anode of the diode is made positive with respect to its cathode. The valve therefore conducts causing electrons to flow from ground through R and from cathode to anode in the valve. This electron flow develops a p.d. across R. The input voltage divides between R and V in the ratio of their resistances. The resistance of V is usually a few hundred ohms when conducting while the resistance of R may be several thousand ohms. Hence the output voltage developed across R has the same waveform as the positive half cycle of applied voltage and an amplitude almost as great.

During the negative half cycle of the input voltage, the anode of the valve is driven negative with respect to its cathode. The valve does not conduct during this time, but acts as an open circuit or an infinite impedance. Since no current flows through R, there is no voltage developed across it during this

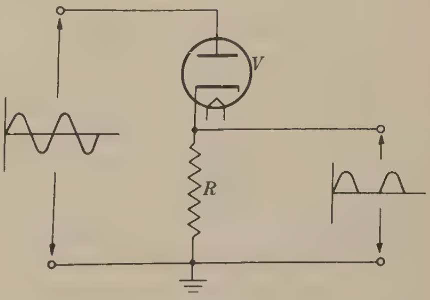


FIG.15-1 Negative-clipping circuit with a series diode.

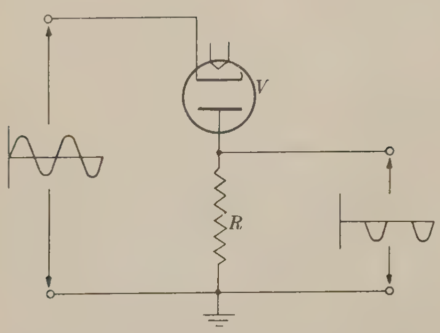


FIG.15-2 Positive-clipping circuit with a series diode.

time. Consequently the negative half of the waveform is clipped off by the limiter. Hence, the output voltage is a series of positive pulses as in Fig. 15-1.

The positive half instead of the negative half of the input voltage may be clipped off by using the circuit shown in Fig. 15-2. In this circuit the diode valve has been inverted. It conducts during the negative half cycle of applied voltage when the cathode is made negative with respect to the anode. Electron flow from cathode to anode, through *R*, develops the output voltage across *R*. During the positive half cycle of input voltage the cathode is positive with respect to the anode. No current flows during this time and there is no voltage developed across *R* until the negative half cycle recurs. The output voltage consists of a series of negative pulses as in Fig. 15-2.

B. Parallel-Diode Clipping

The clipping effect of a diode may also be utilized by connecting the output circuit in parallel with the diode as in Fig. 15-3. As before the resistance of *R* must be large in comparison with that of *V* when the latter is conducting. In addition the impedance of a load circuit connected across the output must be large in comparison with the resistance of *V* and the resistance of *R*. During the positive half cycle of input the anode of the valve is made positive with respect to its cathode. The valve conducts and the current flowing through it develops a p.d. across *R* almost as large as the applied voltage. The remaining small voltage across the diode is applied to the output circuit. When the input voltage becomes negative, the valve stops conducting and acts as an open circuit. The input voltage divides across *R* and the load circuit in the ratio of their impedances. Since the impedance of the load circuit is large in comparison with the resistance of *R*, most of the input voltage appears across the load circuit. The output waveform is shown in Fig. 15-3. The positive half cycles are square and of very small amplitude. The negative half cycles of large amplitude have the shape of the negative half input cycles.

By reversing the connections to the diode (Fig. 15-4) the valve is made to conduct during the negative half cycle of input and presents a low impedance in parallel with the load at this time. During the positive half of the input cycle, the valve does not conduct and presents an infinite impedance in parallel with the load circuit. The output waveform (Fig. 15-4) is the same as that for Fig. 15-3 except that it is inverted.

With this parallel type of circuit the output voltage across the valve is limited to such a small value on the half cycles when the diode conducts that the clipper may be considered to limit the output voltage to zero during one half of the input cycle.

C. Biased-Diode Clipping

It is sometimes required to limit a given voltage to a definite positive or negative value. Fig. 15-5 shows a biased diode that may be used to accomplish this. In this example a sinusoidal voltage of 10 volts amplitude is limited to 6 volts amplitude in the positive direction in the output. A bias battery of 6 volts maintains the cathode of the valve 6

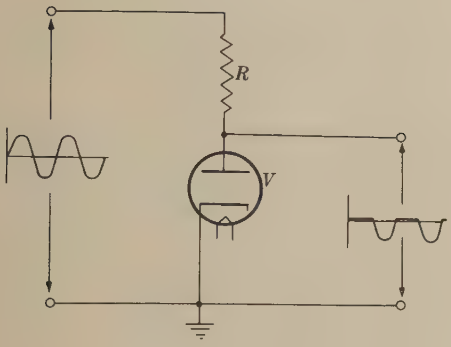


FIG. 15-3 Positive-clipping circuit with a parallel diode.

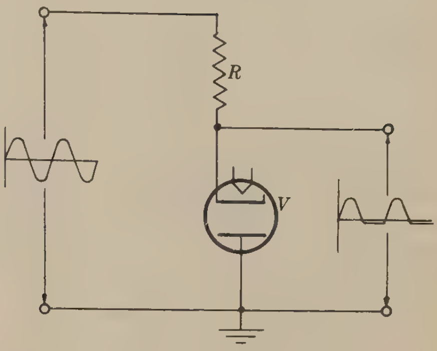


FIG. 15-4 Negative-clipping circuit with a parallel diode.

volts positive to ground and to the anode in the absence of a signal. On the positive half cycle of the input the diode does not conduct until the input rises above 6 volts. As long as the input is less than 6 *V* there is no voltage drop across *R* and the output is the same as the input. As the input voltage increases above 6 *V*, the anode becomes positive to the cathode and the valve conducts and becomes a very low impedance. Any increase in input voltage in excess of 6 volts is developed across *R* so that the output on the positive half cycle is limited to 6 volts, the voltage of the bias battery. On the negative half cycle of the input, the valve does not conduct and no p.d. is developed across *R*. The output waveform therefore follows the input during this part of the cycle. Thus, the output voltage (Fig. 15-5) has a flat-topped positive half cycle of 6 volts amplitude and a negative half cycle of the same wave shape and amplitude as the input. Increasing the bias decreases the width of the flat-topped portion of the wave. If it is desired to limit the amplitude of the output voltage in the negative direction, the diode and battery connections are reversed (Fig. 15-6). On the positive half cycles of the input the diode does not conduct and the output waveform is the same as the input. On the negative half cycles the output waveform is the same as the input until the input exceeds −6 *V*. When the input goes more negative than 6 *V* the diode conducts and the negative clipping results.

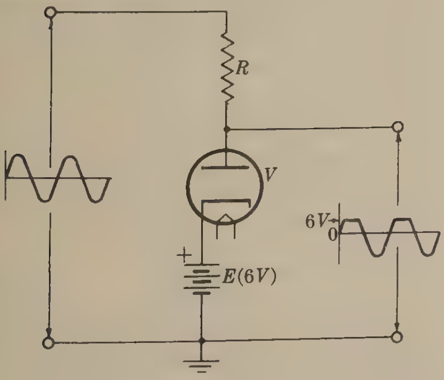


FIG. 15-5 Positive clipping with a biased diode.

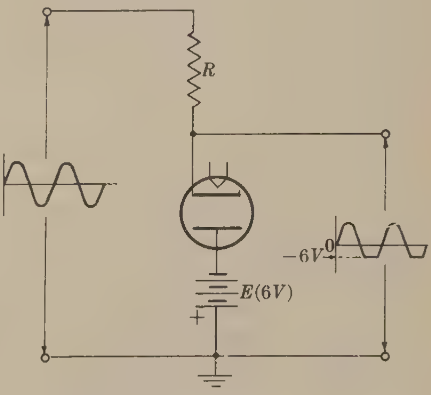


FIG. 15-6 Negative clipping with a biased diode.

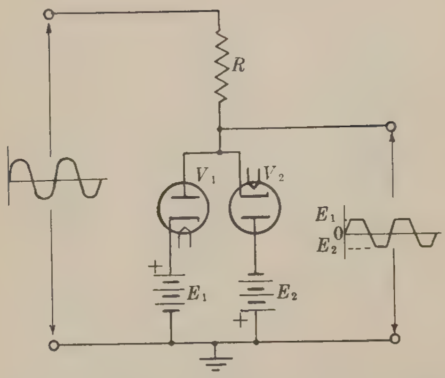


Fig.15-7 Positive and negative clipping with biased diodes.

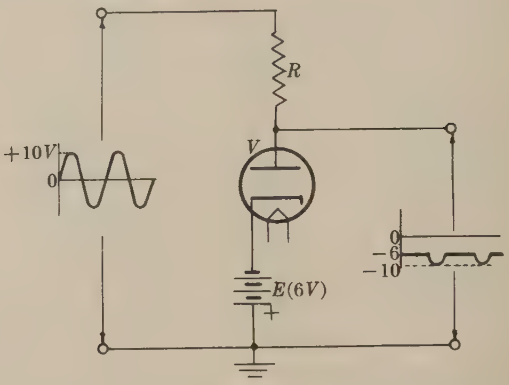


Fig.15-8 Voltage limiting with a biased diode.

Both the positive and negative extremes may be limited by the use of two biased diodes (Fig. 15-7). V_1 and R operate in the manner just described to limit the positive output voltage to that of the bias battery. V_2 and R operate on the negative half of the input cycle, limiting the output voltage to the p.d. of E_2. The output (Fig. 15-7) consists of alternate positive and negative half cycles with flat top waveforms. Such a circuit may be used to produce a substantially square wave from a sine wave.

Another type of parallel-diode limiting is illustrated by the circuit of Fig. 15-8. It is similar in appearance to the limiter of Fig. 15-5 but the bias battery has been reversed making the cathode negative with respect to ground and the anode. When the p.d. between the input terminals is zero, the valve conducts and the p.d. across R is almost equal to the bias voltage since the impedance of the conducting diode is very small compared to that of R. Hence the output voltage equals -6 volts. As the input voltage increases in the positive direction, the valve conducts more heavily and the increase in the voltage across R is practically equal to the increase in the input voltage. Hence the output voltage remains the same, that is $-6V$, as the input voltage increases in the negative direction. The valve continues to conduct as long as the input is not more negative than $-6V$ and the output voltage remains the same at $-6V$. When the input voltage exceeds 6 volts in the negative direction, however, the anode of the valve becomes negative to the cathode. The valve stops conducting and the voltage across R disappears. The remainder of the negative peak of the input waveform in excess of 6 volts is therefore applied to the output terminals. Such a circuit is used to pass only the negative extreme of input voltage to the succeeding stage. The output is a varying negative voltage the negative pulses being of short duration compared to the time between pulses.

By connecting a suitable coupling capacitor and resistor in series across the output terminals an a-c output voltage is obtained (Fig. 15-9). The positive part is nearly zero so that the output is a series of negative pulses. If positive pulses are required, both the valve and the bias battery are reversed as shown in Fig. 15-10.

D. Grid Clipping

The positive portion of an alternating voltage waveform may be clipped off by means of the *grid-current limiter* shown in Fig. 15-11. The input voltage is applied

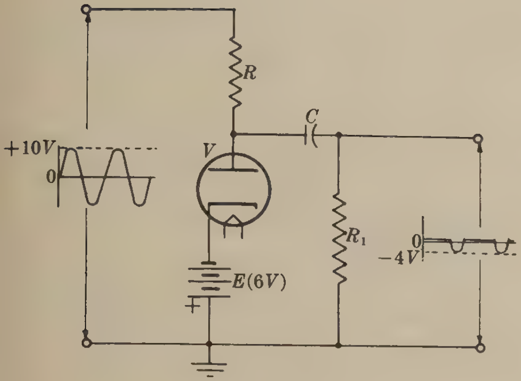


FIG.15-9 Limiting circuit used to produce negative pulses.

through a large series resistor R_1 to the grid of the valve. The cathode is connected to ground so that there is no bias. The load resistor R_2 and battery complete the anode circuit.

The grid-cathode circuit of the triode operates in a manner similar to the anode-cathode circuit of the diode limiters. When the input voltage becomes positive, the grid attracts electrons from the cathode causing grid current to flow through R_1. R_1 is made large in comparison with the grid-cathode resistance of the valve, so that most of the input voltage is developed across R_1. The actual grid voltage is limited to a small positive value. When the input voltage is negative no grid current flows and all the input voltage appears between grid and cathode, no voltage appearing across R_1.

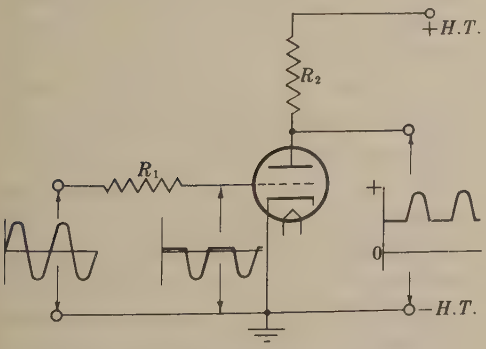


FIG.15-11 Grid-clipping circuit.

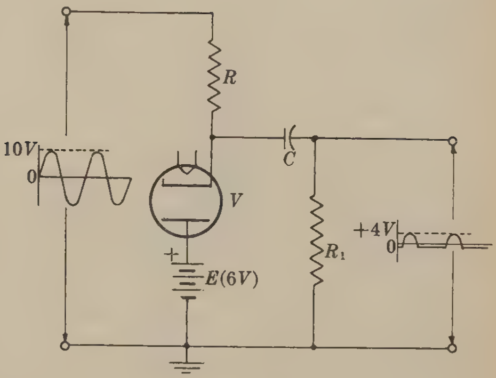


FIG.15-10 Limiting circuit used to produce positive pulses.

During the positive half cycle of input, the anode current is large and remains at a steady value since the grid has a small steady positive potential. There is a large p.d. across R_2 so that the anode potential has a low steady value. When the input voltage becomes negative the grid potential follows the variations of input voltage. The current through the valve decreases, causing a rise in the anode potential. A sine wave of voltage in the input ordinarily produces a series of amplified half sine waves at the anode of the valve as shown in the diagram provided that the input does not drive the grid below cut-off.

The limiting action of the grid circuit may be delayed by biasing the grid negatively (Fig. 15-12) with respect to the cathode. This may be accomplished

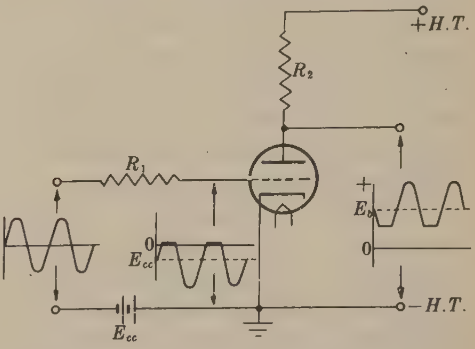


FIG.15-12 Biased grid-clipping circuit.

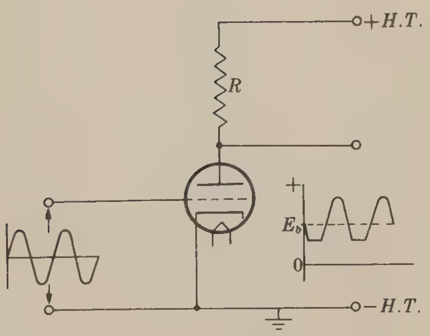


FIG. 15-13 Saturation limiting circuit.

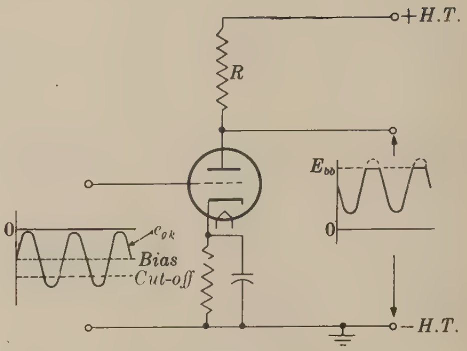


FIG. 15-14 Cut-off limiting circuit.

by a negative bias battery E_{cc} connected in the grid circuit or by a cathode resistor and by-pass capacitor to maintain the cathode positive to ground. The input voltage on the positive half cycle must first overcome the bias voltage and drive the grid positive before clipping takes place. Thus the width of the flat portion of the grid voltage and output voltage waveforms is considerably less than half a cycle.

E. Saturation Limiting

If the grid limiting resistor and bias battery of the previous circuit are omitted as shown in Fig. 15-13, limiting of the positive half of the input cycle may result due to anode current saturation. As the grid is driven positive, the current through the valve increases and the p.d. across the load resistor R increases. The anode potential of the valve therefore decreases. In order that anode current may flow the anode voltage must never fall to zero and thus the maximum voltage across R must always be less than the supply voltage E. Hence, the maximum anode current I must always be less than a value determined by

$$I = \frac{E}{R}$$

where E = p.d. of the supply battery
R = resistance of the load.

If R has a large value, the maximum current need not be large to cause anode current saturation. The saturation condition can exist without having all of the electrons emitted by the cathode attracted to the anode. This condition may be produced by making the grid only a few volts positive to the cathode in a small valve. Any further increase in grid voltage during the positive half of the input cycle does not produce any further change in the anode current and hence in the output voltage. A larger amplitude of output waveform is possible with this circuit than with the grid limiter since the grid is driven more positive and the anode potential falls to a lower value. The circuit requires a larger driving power to maintain the grid positive during part of each cycle. Also the impedance of the driving source must be low to prevent grid clipping.

F. Cut-Off Limiting

Limiting of the amplitude on the negative half of the input cycle may also be accomplished with the circuit of Fig. 15-14. This is most likely to occur when the grid is biased negatively with respect to the cathode, or when a large signal is applied to the grid. If the input voltage drives the grid to cut-off, the p.d. across the load resistor falls to zero and the anode potential rises to the value of the supply voltage. Any increase in negative voltage on the grid does not produce any further change in anode current. Consequently a waveform with a flat positive top on each cycle is produced at the anode.

G. Overdriven Amplifier

An *overdriven amplifier* is a circuit which is used to produce a square wave output from a sine wave input. It combines the operational characteristics of the anode-current-saturation limiter and the cut-off limiter previously described. A large input signal is applied to the grid of a small valve which has a large load resistance R connected in its anode circuit (Fig. 15-15). The valve may be unbiased if an unsymmetrical output is allowable. If equal limiting on positive and negative half cycles is required, the grid is biased midway between the potentials which produce anode current saturation and anode current cut-off.

On the positive half cycle of input voltage, the grid is quickly driven to the anode current saturation point. The valve conducts heavily and a large p.d. is developed across the load resistor. During the time of saturation the anode voltage is low and almost constant. On the negative half cycle of the input voltage, the grid is rapidly driven beyond cut-off. The anode potential rises to the supply voltage and remains at that value until the grid is returned above cut-off.

Successful operation of this circuit requires a driving source with low output impedance to drive the grid positive each cycle. The waveform produced at the anode is practically square as shown in the diagram.

15.2 Clamping circuits. A *clamping circuit* is one which holds either amplitude extreme of an a-c voltage to a definite level of potential without distorting the a-c waveform. In this respect it differs from a clipping circuit which removes a part of the waveform. The clamping circuit operates by changing the d-c component of the voltage waveform and is therefore also known as a *d-c restorer.*

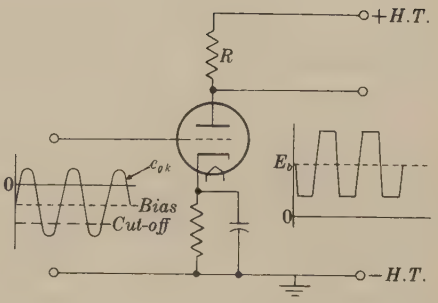


FIG.15-15 Clipping circuit using an overdriven amplifier.

Before proceeding to the description of actual clamping circuits, it is necessary to analyse in some detail the operation of a resistance-capacitance coupling network. Such a network is shown in Fig. 15-16 coupling together two triode amplifiers V_1 and V_2. The voltage used in this consideration is a square wave, one of the waveforms associated with clamping circuits. This square wave is applied to the grid of V_1 producing the square wave of voltage at the anode as shown in the diagram. During the square-wave cycle *a*, *b*, *c*, *d*, *e*, the anode potential of V_1 undergoes the following changes. The anode voltage of V_1 remains steady at a minimum value of 20 volts from *a* to *b*, a period of 35 microseconds. It then rises abruptly to 100 volts at *c* and remains steady for the next 5 microseconds from *c* to *d*. At *d* the anode voltage falls to 20 volts at *e* and remains constant again during the first part of the next cycle.

The coupling capacitor C and the grid-leak resistor R are chosen to have a time constant which is large in comparison with the 40 microsecond period of the cycle. Also, the reactance of C must be small in comparison with the resistance of R at the fundamental frequency, 25 Kc/s, and harmonic frequencies which are included in the rectangular wave-

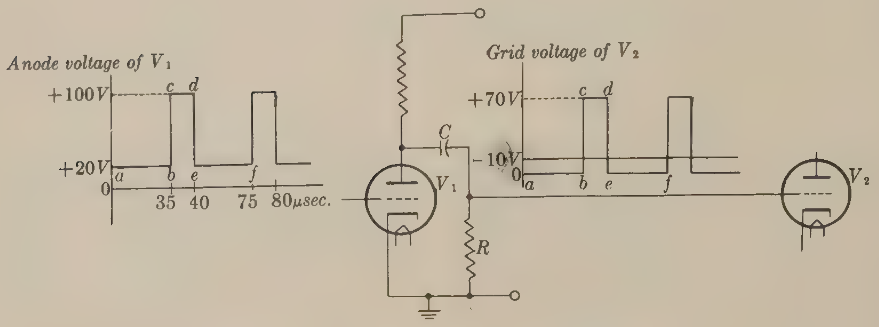


FIG. 15-16 R-C coupling of square-wave voltage.

form. Under these conditions R and C may be considered to form an ideal coupling network between V_1 and V_2.

When the anode voltage of V_1 is steady at 20 volts at b the capacitor is charged to 20 volts plus an additional 10 volts due to the more positive portion of the preceding cycle. At this time the voltage across C is 30 volts and C is discharging through R so that the grid voltage of V_2 is -10 volts. When the anode potential of V_1 rises to 100 volts from b to c an 80 volt increase is applied to C and R in series. The p.d. across C cannot change instantaneously; consequently this rise of voltage is applied to R. That is, at this instant C stops discharging and begins to charge through R and produces this 80 volt change across R. The grid voltage of V_2 changes from -10 volts to $+70$ volts, from b to c. Since the time constant of C and R is long in comparison with the period of the applied cycle, the p.d.'s across C and R do not change greatly during the period cd. From d to e, the anode potential of V_1 falls suddenly to 20 volts. C stops charging and begins to discharge because the anode potential is less than the p.d. across C. The discharge of C through R makes the upper end of R 10 volts negative with respect to ground.

The voltage applied to the grid of V_2, which is the p.d. across R at every instant, is therefore an alternating voltage. It is positive when the anode potential of V_1 is high and negative when the anode potential of V_1 is low. Moreover, the capacitor must receive the same amount of charge during the positive portion of the cycle as it loses during the negative portion. Therefore the area on the positive side of the V_2 grid voltage waveform equals the area on the negative side. This condition is fulfilled if the grid potential is 70 volts positive for 5 microseconds and 10 volts negative for 35 microseconds during each cycle. In practice, of course, it is impossible to maintain perfectly square waves of voltage. The potential at the anode of V_1 is prevented from rising and falling instantaneously by the interelectrode capacitances of the valve. This effective capacitor must be charged or discharged each time the anode potential changes. The rate at which the anode of V_1 is capable of changing p.d. depends upon the resistance in the circuit as well as the value of capacitance. Moreover, the time constant of the R-C coupling network can never be made infinite. Consequently the flat portions of the grid voltage waveforms are actually sections of exponential curves which tend always to approach the zero axis as shown in Fig. 15-17.

In the preceding resistance-capacitance coupling circuit the output applied to the grid of V_2 is an a-c voltage which

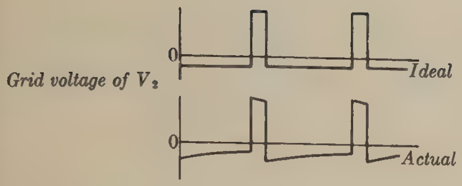


Fig. 15-17 Ideal and actual waveforms of grid voltage of V_2.

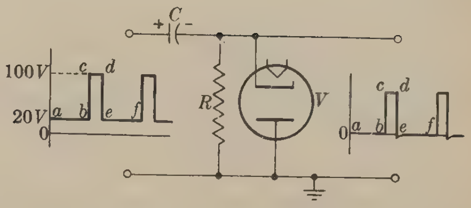


Fig. 15-19 Diode clamping circuit producing a voltage clamped to positive side of ground potential.

alternates about ground potential. The coupling capacitor blocks the d-c component from the grid circuit. In pulsed circuits it is often necessary that the output voltage of the R-C coupling circuit swing entirely positive or entirely negative with respect to ground potential (Fig. 15-18). That is a d-c component must be restored to the grid voltage. The voltage waveform is effectively clamped to one side of ground potential by a *clamping circuit* or *d-c restorer*.

A. Diode Clamping

A diode circuit which may be used to clamp the output waveform of an R-C coupling circuit to the positive side of ground potential is shown in Fig. 15-19. A diode valve V is connected in parallel with the resistor R in the coupling network. In this example, the input voltage to the R-C circuit is a square wave voltage, the same as that in the previous example.

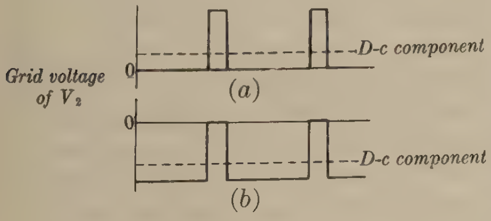


Fig. 15-18 Voltage clamping (a) Waveform clamped to positive side of ground potential (b) Waveform clamped to negative side of ground potential.

When the input voltage is steady from a to b at 20 volts the coupling capacitor is charged to 20 volts also, due to discharge through the diode. Since C is neither charging nor discharging, there is no flow of electrons through R and consequently no p.d. across it. Therefore the voltage applied between anode and cathode of the diode is zero and the diode does not conduct. The output voltage is zero from a to b.

When the input voltage rises from b to c, the coupling capacitor begins to charge slightly through R developing the positive rise bc in the output. The cathode of V is positive with respect to ground and the anode and so the diode does not conduct. It is an infinite impedance in parallel with the output and has no effect on the output waveform. Since the time constant of C and R is long in comparison with the period of the applied cycle, the p.d.'s across C and R do not change greatly during the period cd.

From d to e the input voltage falls sharply to its previous level of 20 volts. The p.d. across the capacitor is slightly more than 20 volts and so C discharges making the upper end of R slightly negative to ground and the cathode of V negative to its anode. The diode therefore conducts and becomes a low resistance in parallel with R. C discharges much more quickly through V than it would through R because of the reduced time constant. This short discharge time is indicated by the brief negative potential at e in the output waveform. When the p.d. across C has become steady at 20 volts, the p.d. across R and V disappears and the circuit is ready to

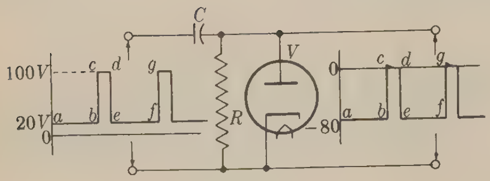


Fig.15-20 Diode clamping circuit producing a voltage clamped to negative side of ground potential.

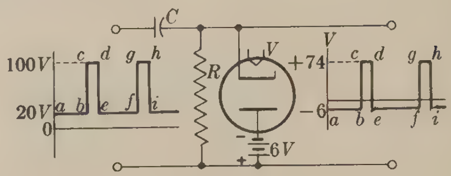


Fig.15-21 Clamping circuit using positively biased diode.

repeat its cycle as before. The diode permits the use of a large coupling resistor *R*. This provides a very long time constant during the portion *cd* of the input cycle and the diode provides an extremely short time constant during the portion *ab*.

The output voltage may be clamped to the negative side of ground potential by reversing the diode as in Fig. 15-20. In this case it is convenient to begin the description at a time *c* when the input voltage reaches its maximum of 100 volts. The capacitor *C* charges rapidly to 100 volts mainly through the low resistance of the valve which conducts at this time. A very tiny part of the charging current flows through *R*. The small positive tip of output voltage at *c*, produced by the charging current through *R* and *V* indicates the charging period of the capacitor. From *c* to *d* when the input voltage is steady at 100*V* and the capacitor is charged to 100*V* there is no current through *R* or *V* and the output voltage is zero. When the input voltage falls to 20*V* at *e*, a p.d. of 80 volts is developed across *R*, due to the discharge of *C*, and the upper end of *R* is negative. *C* continues to discharge to maintain this p.d. during the period *ef* due to the long time constant of *C* and *R*. The valve is inoperative during this time as its anode is negative with respect to its cathode. From *f* to *g* the input voltage rises to 100*V*. The capacitor stops discharging at the point *g* and recharges quickly through the valve to complete the cycle. While the description covers the case of an unsymmetrical square wave, it is equally applicable to voltage waveforms of a wide range of shapes common in pulsed circuits.

B. Biased Clamping

The voltage waveform may be clamped to any reference voltage other than zero by means of a biased-diode clamping circuit. Fig. 15-21 shows a circuit that may be used to clamp the input voltage to the positive side of −6*V* without distorting the waveform. The anode of the diode is biased 6 volts negative to its cathode by the battery. Thus *V* does not conduct until the negative voltage across *R* exceeds 6 volts. At *b* in the voltage waveforms the input is 20*V*, the voltage across the capacitor is 26*V*. Thus, the capacitor is discharging through *R* producing an output voltage across *R* of −6*V*. The input rises from 20*V* to 100*V* from *b* to *c*. *C* stops discharging and begins to charge through *R* producing a change of 80*V* across *R*. The output voltage changes from −6*V* to +74*V*. From *c* to *d* the diode does not conduct and since the time constant of *C* and *R* is long in comparison with the period of the applied cycle the p.d.'s across *C* and *R* do not change greatly. From *d* to *e* the input voltage falls suddenly to 20*V*. *C* which charged to slightly more than 26*V* stops charging and begins to discharge and the voltage across *R* becomes slightly more than −6*V*. The cathode of the diode goes slightly negative to the anode and the diode conducts. *C* quickly discharges

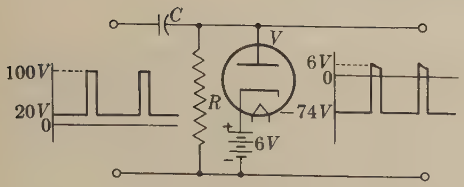


FIG.15-22 Clamping circuit using negatively biased diode.

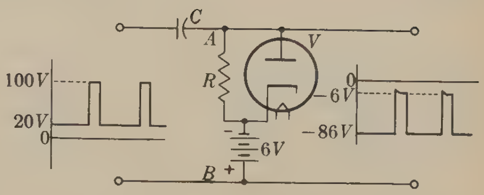


FIG.15-23 Clamping circuit using negatively biased coupling network.

through the diode and its voltage quickly drops to 26*V*. The rapid discharge through the valve is indicated by the negative pip at *e*. From *e* to *f*, *c* continues to discharge through *R* producing a voltage across *R* of −6*V*, which is steady because the time constant of *C* and *R* is long.

By reversing the diode and battery connections the circuit may be made to clamp the waveform to a limit of 6 volts in the positive direction (Fig. 15-22). When the voltage across *R* is less positive than +6*V* or is negative the diode does not conduct. When the voltage across *R* goes more positive than +6*V* the diode conducts and quickly charges the capacitor. The output voltage is thus prevented from rising above +6*V*.

Clamping below ground potential may be accomplished by biasing the entire coupling network negatively as shown in Fig. 15-23. If the voltage between *A* and *B* is more negative than −6*V*, the anode of the valve is negative with respect to its cathode and the valve does not conduct. When the voltage between *A* and *B* rises to −6*V* the voltage across the valve becomes zero as the cathode is maintained at −6*V* by the bias battery. If the voltage between *A* and *B* is less negative than −6*V*, the valve conducts and charges the capacitor. The output voltage is therefore prevented from rising above −6*V*.

While the variations in clamping circuits have not been exhausted in the foregoing examples, the basic types of diode clampers have been outlined. The operation of other clamping circuits may be analysed in a similar manner.

C. Grid Clamping

The grid-cathode circuit of most valves may perform the function of a diode clamping circuit. Fig. 15-24 illustrates a circuit in which the input waveform is confined to operate on the negative side of ground potential at the grid of the valve. When the input voltage rises, the grid is driven positive to ground and hence to the cathode of the valve. Grid current flows to charge the coupling capacitor quickly through the comparatively low resistance of the grid-cathode circuit. Hence the grid voltage is limited to a very small value in the positive direction.

When the input voltage falls however, the coupling capacitor discharges through *R* only since the grid is negative to the cathode. This action develops the full input voltage at the grid on the negative side of ground potential. Thus the output voltage from the amplifier has the same waveform as the input except that it is inverted and amplified.

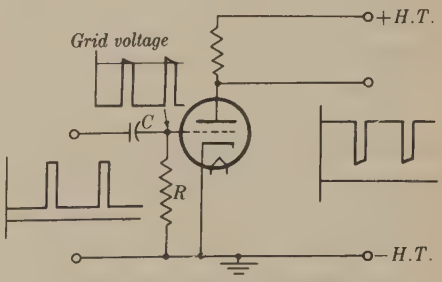
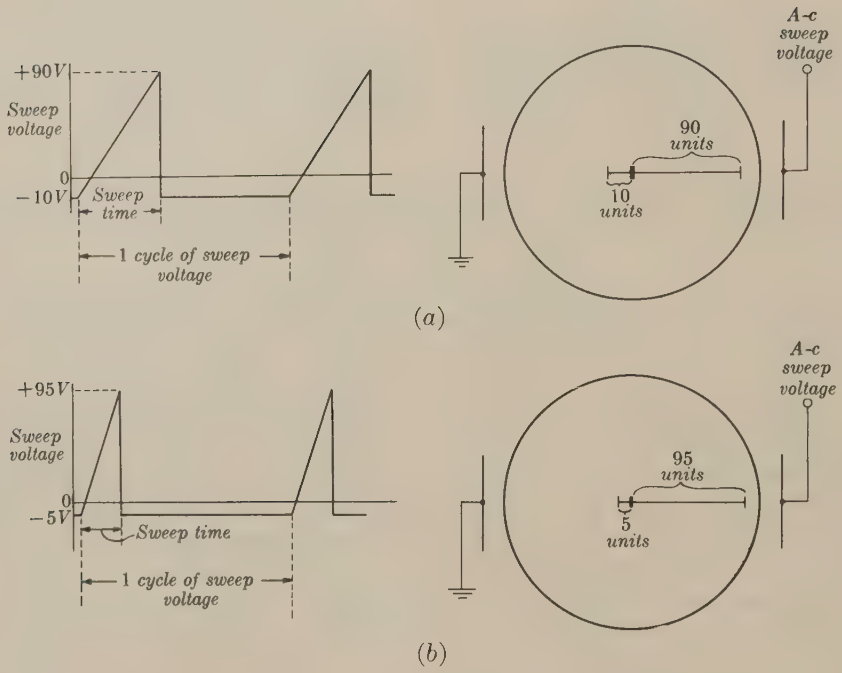


FIG.15-24 Grid clamping circuit.



(a)

(b)

FIG. 15-25 Shift of trace on c.r.t. due to change in position of range switch.

D. Application of Clamping Circuits

Clamping circuits are widely employed in the sweep circuits of cathode-ray tubes. The necessity for such circuits is illustrated in Fig. 15-25 (a) and (b) which shows two a-c sweep voltages of the same amplitude and same frequency but with different sweep times which are to be applied in turn to the right *X*-plate of a c.r.t. Both voltages are indicated as having a variation of 100 volts from base to peak. The a-c voltage of Fig. 15-25 (a), which swings 10 volts on the negative side and 90 volts on the positive side of its reference voltage causes the trace to be positioned as seen in the figure.

If the waveform of Fig. 15-25 (b) is applied to the right *X*-plate, the trace is observed to be shifted to the right. The portion of the sweep voltage during the sweep time is narrower in this case, causing the voltage to rise to 95 volts in the positive direction and fall to −5 volts in the negative direction. At the start and at the end of the sweep, the right *X*-plate is 5 volts more positive in Fig. 15-25 (b) than in (a). This causes the trace to shift sideways when the change is made from one sweep voltage to the other. In order to avoid this inconvenience the sweep voltage is applied to the deflection plate via a clamping circuit (Fig. 15-26). The clamping circuit ensures that each sweep voltage operates the full 100 volts on the positive side of the ground reference level and does not swing in the negative direction. The trace is therefore made to begin at the same point on the screen regardless of the input waveform. In actual practice the horizontal centering controls set the position of this point so that the trace is properly centered. This is especially important in cathode-ray tubes used in radar sets in which the trace must remain aligned with a scale fixed on the outside of the tube.

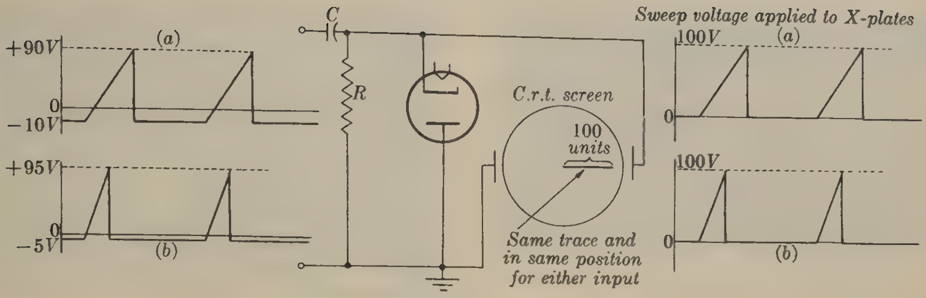

+90V
(a)
0
−10V
+95V
0
−5V
(b)
C
R
C.r.t. screen
100 units
Same trace and in same position for either input
Sweep voltage applied to X-plates
100V
(a)
0
100V
0
(b)

FIG. 15-26 Clamping circuit used to prevent horizontal trace shift in unbalanced c.r.t. sweep circuit.

While the explanation is given for an unbalanced c.r.t. circuit above, the necessity for clamping circuits is equally important in balanced circuits. The method of connecting diode clamping circuits in a balanced c.r.t. sweep circuit is shown in Fig. 15-27. The a-c sweep voltages are applied to the horizontal-deflection plates via the clamping circuits. The diodes V_1 and V_2 are connected in parallel with the coupling resistors R_1 and R_2 to ensure that the trace starts at the same point for different sweep voltages. The horizontal centering control (not shown) is adjusted as before to set the position of the starting point of the trace.

Clamping circuits are equally necessary in the sweep current circuits of cathode-ray tubes which employ magnetic deflection. Further applications of clamping circuits are found in gating circuits which are discussed in later sections.

The voltage waveforms shown in the diagrams are somewhat idealized in order to emphasize the points under discussion. It is equally important that the technician be able to interpret voltage waveforms obtained by monitoring the circuit with a cathode-ray oscilloscope. For this reason, oscillograms showing the operation of actual circuits have been included periodically throughout the present chapter.

Fig. 15-28 shows oscillograms of the sweep voltage monitored at the input and output of the sweep amplifier. The first shows the grid voltage and the second the anode voltage of the amplifier producing a 20-mile sweep for a surface-warning set.

The operation of a clamping circuit is shown by the oscillograms of Fig.

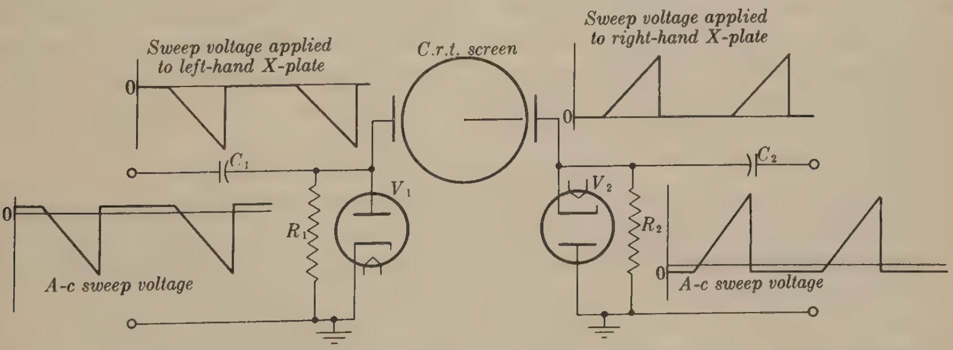
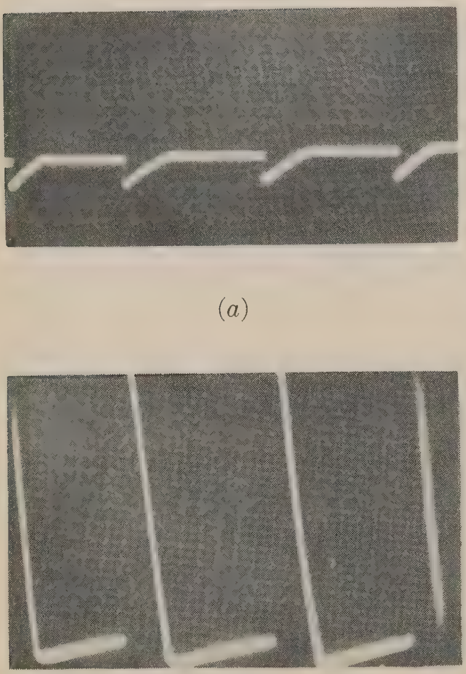

Sweep voltage applied to left-hand X-plate
0
C_1
0
A-c sweep voltage
R_1
V_1
C.r.t. screen
Sweep voltage applied to right-hand X-plate
0
C_2
V_2
R_2
0
A-c sweep voltage

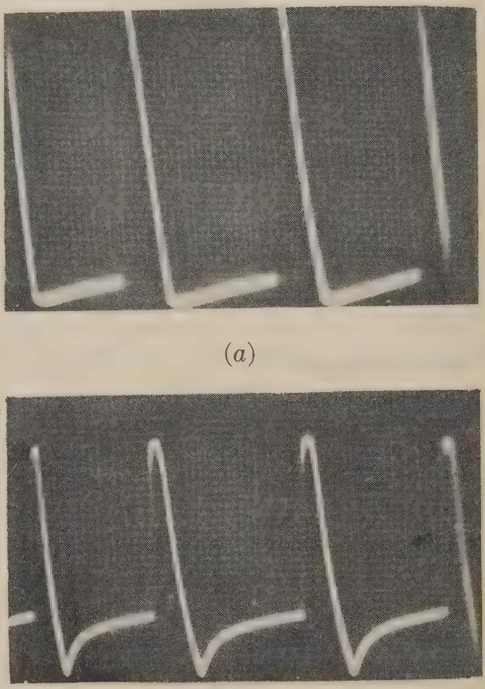
FIG. 15-27 Clamping circuit used to prevent horizontal trace shift in balanced c.r.t. sweep circuit.



(a)

(b)

Fig. 15-28 Oscillograms of sweep voltage waveforms.



(a)

(b)

Fig. 15-29 Oscillograms to show the operation of a clamping diode on sweep voltage waveforms (a) Clamping diode removed (b) Clamping diode in operation.

15-29. The first shows a sweep voltage waveform developed across the coupling resistor to the deflection plate of the c.r.t. when the clamping diode is removed. The second shows the same voltage with the clamping diode in operation.

15.3 Pulse forming circuits. The accuracy of the many timing circuits used in radar and television depends in large measure upon the development and application of precise trigger pulses. A trigger pulse waveform is one of a number of pulse waveforms. Several common types of pulses are shown in Fig. 15-30. For each pulse waveform the duration of the pulse is short compared to the interval between pulses. The number of pulses per second, which is the same as the number of cycles per second in the waveforms illustrated is called the *pulse repetition rate* or *pulse recurrence rate*. The kind of pulse shown in Fig. 15-30 (d), the duration of which is extremely short compared to the interval between pulses, is called a *trigger pulse*. A circuit whose operation is initiated by such a pulse is called a *triggered circuit*. The limitations placed upon these pulses varies considerably with different applications, but three characteristics are usually of prime importance. These are

(i) the duration of the pulse
(ii) the waveform
(iii) the repetition rate.

Frequently the duration of the trigger pulse must not be greater than a few microseconds and may even be a fraction of one microsecond.

The waveshape of the pulse must be precise in order that a definite amplitude be attained at a definite instant to

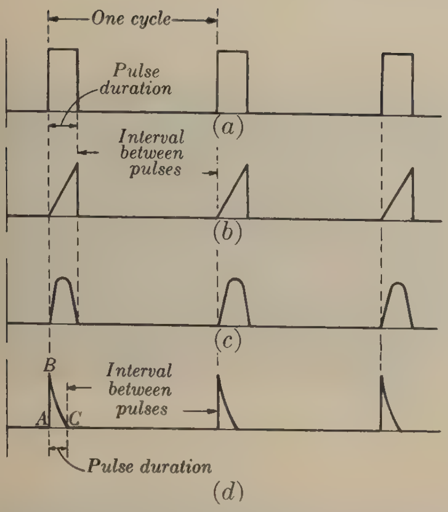


FIG.15-30 Common types of pulse waveforms.

initiate triggering. This ordinarily requires that the leading edge AB of the trigger pulse be as steep as possible. The trigger-pulse amplitude, for example, may change by $100\,V$ in a fraction of a microsecond. The trailing edge BC should be fairly steep also since, if the pulse were not quickly removed, the triggered circuit might respond unfavourably to the pulse. A second cycle of operation might be started by a single pulse, or, the trailing edge of the trigger pulse might begin an undesired sequence of operations. For some applications the amplitude of the pulse must be accurately controlled and must not be altered by changes in other circuits.

Finally the trigger pulse must ordinarily be made to recur at a precise time at repetition rates that vary from a few cycles per second to a hundred thousand cycles per second. Any variation in the time interval between successive pulses introduces inaccuracies into the performance of the triggered circuits.

Two examples of the specifications of radar pulses will serve to emphasize the precision required of pulse forming circuits.

(i) A calibration circuit must supply to the Y-plates of the c.r.t. a pulse of 50 volts amplitude and 1 microsecond duration every 10 microseconds, or 100,000 times per second.

(ii) A modulator must supply a uniform negative voltage pulse of 10 kilovolts amplitude at a current of 10 amperes lasting for 1.5 microseconds and recurring every 2000 microseconds.

A. *Differentiation*

The circuit of Fig. 15-31 is frequently used to develop a series of sharp positive and negative pulses of output voltage from a square wave of voltage applied to the input terminals. Such a circuit is called a *differentiating circuit* or an *R-C differentiator* and the process is called *differentiation*. The circuit resembles the conventional R-C coupling network but the capacitance of C is small. This provides a very short time constant for R and C. The input is a typical square wave of voltage that rises abruptly from a minimum of 20 volts to a maximum of 100 volts and falls sharply 100 microseconds later.

During the period ab the applied voltage is constant at 20 volts. The capacitor

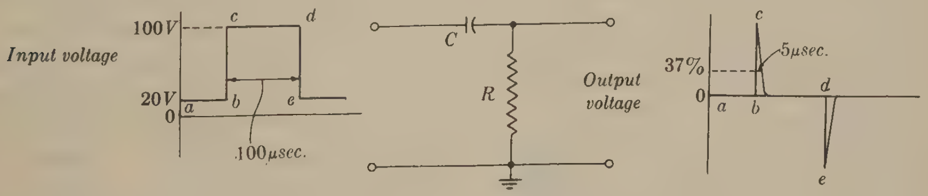
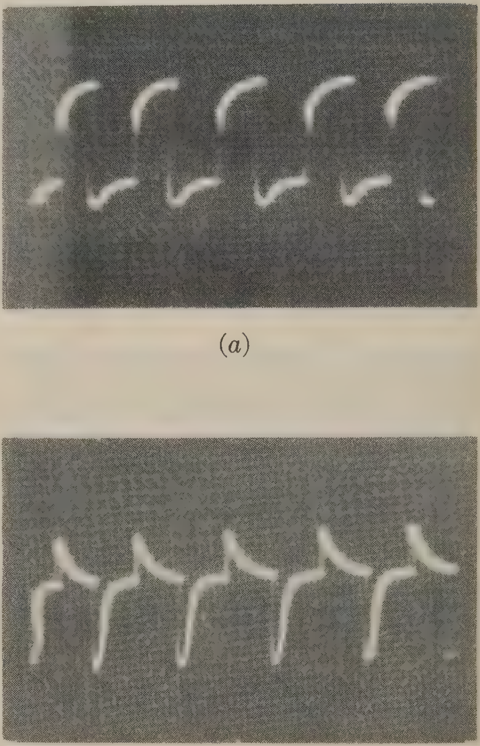


FIG.15-31 R-C differentiating circuit.

is charged to 20 volts and is neither charging nor discharging. There is therefore no flow of electrons through *R* and the p.d. across it is zero as indicated in the output voltage waveform. When the input voltage rises from *b* to *c*, an increase of 80 volts is applied across *R* and *C* in series. Since *C* cannot charge instantaneously, its p.d. cannot change at the first instant. Hence all the increased voltage appears across *R*. *C* begins to charge. The flow of electrons up through *R* provides the p.d. across *R*. The rate at which *C* charges is determined by the time constant of *C* and *R* together with the resistance of the input voltage generating circuit which is here assumed to be negligible. This time constant must be very short compared to 100 microseconds. As *C* charges, the p.d. across *R* decreases exponentially producing the trailing edge of the positive pulse in the output. The p.d. across *R* becomes zero when *C* is charged to 100*V*. The capacitor remains charged to 100 volts during the remainder of the interval *cd* and the output voltage remains at zero during this time.

At *d* the input voltage falls 80 volts to *e*. Capacitor *C* begins to discharge causing electrons to flow down through *R*, developing a p.d. across *R*. At the first instant from *d* to *e*, the p.d. across *c* is 100 volts and the p.d. between the input terminals is 20 volts. Therefore the p.d. across *R* must be −80 volts. This forms the leading edge of the negative pulse of the output waveform. As *C* discharges through the resistance of *R* and the input circuit, the voltage across *R* decreases exponentially to form the trailing edge of the negative output pulse becoming zero when the capacitor voltage is 20*V*.

In practice the input circuit contains capacitances which must be charged or discharged each time there is a change in voltage supplied to the differentiating circuit. For this reason the leading and trailing edges of the input waveform are never ideally vertical.


(*a*)

(*b*)

Fig. 15-32 Oscillograms to show the operation of a differentiating circuit (a) Input voltage waveform (b) Output voltage waveform.

Consequently the leading edges of the output pulses are not exactly vertical as the p.d. across *R* cannot build up faster than the input voltage changes.

Typical values for the differentiating circuit of Fig. 15-29 are $R = 100{,}000$ ohms and $C = 50$ micro-microfarads. This combination has a time constant of $100{,}000 \times 50 \times 10^{-12}$ seconds or 5 microseconds. That is, the output voltage falls to 37% of its initial value of 100 volts in a period of 5 microseconds as indicated in the diagram.

Pulse forming by means of a differentiating circuit requires

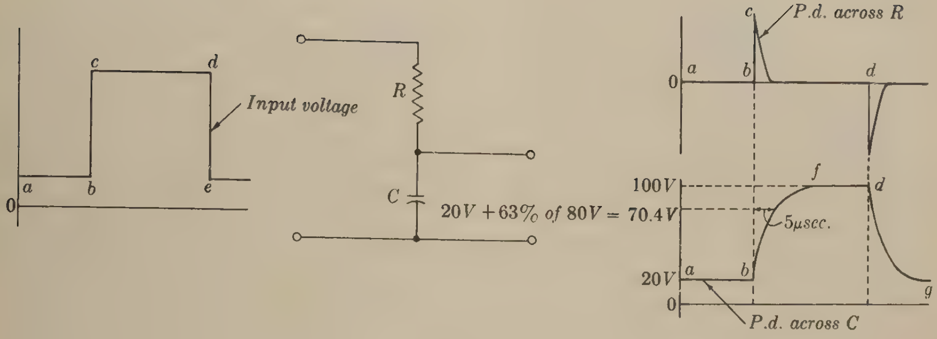


FIG. 15-33 R-C integrating circuit.

(i) that the input waveform have reasonably steep edges

(ii) that the input circuit be capable of maintaining the square waveform in spite of the loading effect of the differentiating circuit. Oscillograms of the input and output voltages of a typical differentiating circuit are shown in Fig. 15-32.

B. Integration

Integrating circuits are not often used as pulse formers but are included at this point because they are similar in operation to differentiating circuits. The circuit of Fig. 15-33 consists of the same components as the differentiating circuit of Fig. 15-31 but the output is taken from the capacitor instead of the resistor. The p.d. across the capacitor at every instant is the difference between the input voltage and the p.d. across the resistor. When the input voltage is 20 volts from a to b, the p.d. across R is zero. Hence the p.d. across C is 20 volts as indicated. When the input voltage rises from b to c, the capacitor charges exponentially from b to f at a rate determined by the time constant of the circuit. This rise in voltage across C from $20\,V$ to $100\,V$ occurs as the voltage across R decreases from its maximum value to zero. When the applied voltage falls from d to c, the capacitor begins to discharge from d to g exponentially. This gradual decrease in voltage across C causes the negative pulse across R.

An integrating circuit is used to delay the leading edge of the input square wave in the output circuit. For example, if R = 100,000 ohms and C = 50 micromicrofarads in this circuit as in the circuit of Fig. 15-31, the time constant is again 5 microseconds. Therefore the p.d. across C rises 63% of its 80 volt change in 5 microseconds. The leading edge of the integrated waveform is delayed by 5 microseconds at an amplitude of $20+\frac{63}{100}$ of 80, or 70.4 volts. If the following circuit is biased so that it starts operating at 70.4 volts input it will be triggered by the voltage across C at a time equal to 5 microseconds after the leading edge of the voltage rise applied to the integrator.

C. Shocked Oscillators

A *shocked oscillator* is a useful circuit for producing oscillations at a definite frequency, beginning at a certain instant and recurring at regular intervals. Reference to a particular application of the circuit will serve to emphasize its usefulness.

The scan of the c.r.t. in a radar set is to be calibrated by a series of pulses spaced 6.1 microseconds apart as indicated in Fig. 15-34. This is done by applying to the upper Y-plate of the tube a sharp positive pulse every 6.1

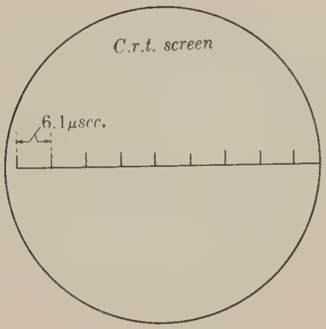


Fig. 15-34 Calibration sweep on a radar c.r.t. screen.

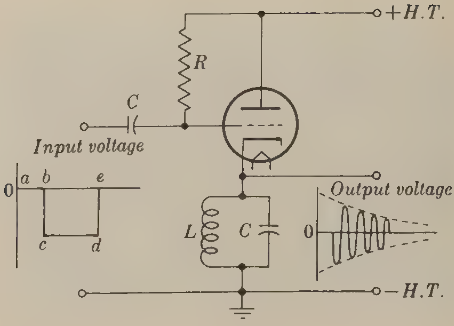


Fig. 15-35 Shocked oscillator for producing a train of oscillations.

microseconds as the sweep progresses from left to right. The first pulse must be synchronized with the start of the sweep to prevent the pattern from drifting and the series must be applied each time the sweep takes place. The interval 6.1 microseconds is commonly used because it is the time required for a transmitted radar pulse or echo to travel 2000 yards or to travel 1000 yards to a target and return. The train of pulses applied to the upper Y plate is produced by an oscillator which operates at a frequency of 163,900 cycles per second. It therefore produces one cycle in $\frac{1}{163,900}$ of a second or 6.1 μsec. Each cycle of the oscillator is used to develop one pulse for the vertical-deflection plates. The pulses are synchronized with the sweep by turning the calibration oscillator on at the start of each sweep and allowing the oscillator to operate for the duration of the sweep. During the waiting period between sweeps, the calibration oscillator is prevented from oscillating. A shocked oscillator is one type of circuit used to produce the calibration pulses for the c.r.t.

A shocked oscillator is shown in Fig. 15-35. A high-Q oscillatory circuit LC is connected between cathode and ground of the triode valve. The resonant frequency of this circuit is set at 163.9 Kc/s by adjustment of L or C. The anode of the valve is connected to the positive side of the supply voltage and the grid is also connected to this voltage point through a large resistor R. With no input signal on the grid, the valve normally conducts heavily. The grid is positive to the cathode causing grid current to flow. This grid current flowing through R develops a large p.d. across R which limits the grid potential to about 1 volt positive to the cathode. The large direct current through the valve flows through L and creates a relatively strong magnetic field around it.

A large negative square wave pulse called a *gate pulse* is applied to the grid of the valve when the oscillations are to commence. This is indicated by the region *bc* of the input waveform. The grid is driven beyond cut-off and the valve acts as an open circuit between the oscillatory circuit and the power supply.

Although current through the valve ceases abruptly, the current through L cannot fall to zero instantaneously. The collapsing magnetic field about L induces a voltage in it which tends to maintain current flowing in the same direction as that at the instant of cut-off. That is, electrons continue to flow from ground to the upper end of L and into the top plate of the capacitor. C is therefore charged negatively on the upper side causing the first half cycle of output voltage to be of negative polarity. A damped oscillatory current continues to flow in the LC circuit until the gate pulse ends. An exponential decrease in the output voltage is caused by losses from the LC circuit in the form of heat, radiation, etc. These are kept to a low value so that oscillations are maintained with relatively constant amplitude for the duration of the gate pulse.

When the gate pulse ends, as indicated by *de*, the grid potential is raised above cut-off and the valve conducts again. The valve and power supply constitute a low impedance load on the oscillatory circuit which quickly damps out any further oscillations. The train of damped sine waves is fed to pulse shaping circuits before being applied to the c.r.t. deflection plates.

The gate pulse applied to the grid of the shocked-oscillator valve must be of sufficient amplitude to maintain the grid beyond cut-off in spite of two factors which tend to make it conduct again. The large oscillation set up in LC makes the cathode considerably negative to ground during one half cycle of operation. This effectively raises the p.d. between cathode and anode of the valve to a value somewhat higher than the supply voltage. In addition, a negative voltage on the cathode is equivalent to a positive voltage on the grid, which further tends to make the valve conduct. Conduction of the valve during the period of oscillation would load the LC circuit and damp out the oscillations prematurely.

A shocked oscillator which responds to only the leading and trailing edges of a square gate voltage on the grid is shown in Fig. 15-36. This circuit is similar to that of Fig. 15-35 except that the oscillatory circuit LC is connected in the anode circuit of the valve. A resistor R is connected in parallel with L and C. R is chosen of such a size that it produces critical damping of the LC circuit. Any oscillation set up in LC is damped out after the first half cycle by R.

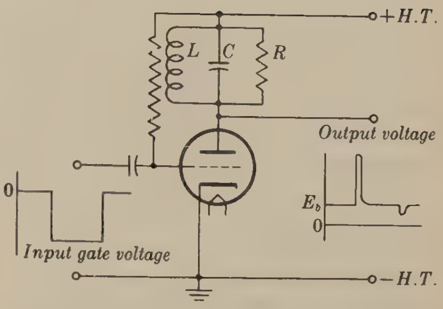


FIG.15-36 Shocked oscillator for producing narrow pulses.

The valve normally conducts heavily as explained for the previous circuit. Since there is little p.d. across L the anode potential is practically the H.T. potential E_{bb}. When the negative field about L collapses one half cycle of oscillation is produced in LC. Since the collapsing field induces a current in L in the same direction as the valve current, the capacitor is charged negatively at the top and positively at the bottom. The oscillation is therefore a positive pulse of voltage from anode to ground. Oscillation ceases after this pulse due to the damping effect of R. At the end of the gate voltage, the valve is made to conduct again. A negative pulse is produced by the LC circuit but its amplitude is less than the positive pulse due to the loading effect of the valve.

For each gate pulse there are two output pulses, one positive and one negative. Thus, the timing of output pulses is controlled solely by the input voltage to the circuit. The exact resonant frequency of LC only affects the duration or width of the pulse produced and so is not critical. For this reason, C is usually omitted from the circuit and the valve capacitance together with the distributed capacitance of the inductor serve to determine the resonant frequency of the

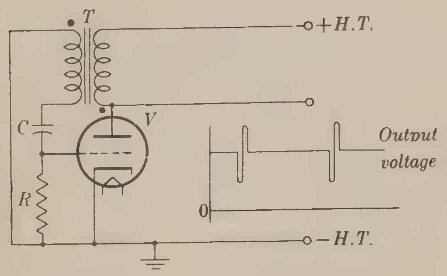


FIG. 15-37 Free-running blocking oscillator.

oscillation. The resonant frequency must be high to produce a narrow pulse.

D. Blocking Oscillators

A *blocking oscillator* is an oscillator which produces one cycle of oscillation and then remains in a quiescent state for a definite period of time. Three types of blocking oscillators are discussed which differ from each other in the manner in which the duration of the quiescent period is determined.

1. FREE-RUNNING BLOCKING OSCILLATOR. The blocking oscillator shown in Fig. 15-37 consists of a triode valve with the primary of a pulse transformer *T* connected in its anode circuit. The secondary of this transformer is connected between grid and cathode of the valve. The coupling capacitor *C* and the grid resistor *R* complete the grid circuit.

When the filament is heated and the high voltage supply is connected the current through the valve increases from zero. This rising current produces an increasing p.d. across the primary of *T* and causes the anode potential to fall (Fig. 15-38). The changing anode potential is coupled through the transformer to the grid of the valve. The secondary winding is so connected that the falling anode potential produces a rising grid potential. The grid is driven positive by the voltage from *T*, causing a further increase in anode current. This process is cumulative and due to the close coupling of *T*, the valve is very quickly driven toward anode current saturation. The grid current charges *C* to almost the full voltage induced in the secondary of *T*. The dip in the anode voltage curve and the peak of the grid voltage curve represent the conditions when the anode current reaches a maximum.

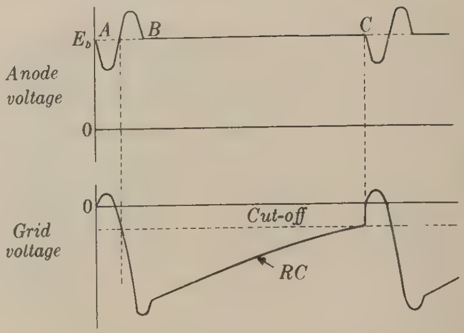


FIG. 15-38 Anode and grid voltage waveforms for a free-running blocking oscillator.

As the anode current approaches saturation, the back e.m.f. across the transformer primary decreases. The anode potential commences to rise again. This is coupled through *T* as a negative voltage applied across *C* and *R* in series. Since the p.d. across *C* cannot change instantaneously, a negative voltage across *R* is applied to the grid of the valve. This falling grid voltage reduces the current through the valve causing the anode voltage to rise sharply. Again the action is cumulative but this time it is in the reverse direction. Anode current is quickly cut off by the large negative potential on the grid. This condition is indicated by the peak of the anode voltage curve and the dip in the grid voltage curve.

The negative grid potential is maintained by the discharge of *C*, electron flow being down through *R*. The time taken for the grid potential to complete its exponential return to cut-off depends upon the time constant of *R* and *C*. When the grid potential reaches the cut-off voltage, current again commences to flow through the valve and the cycle is

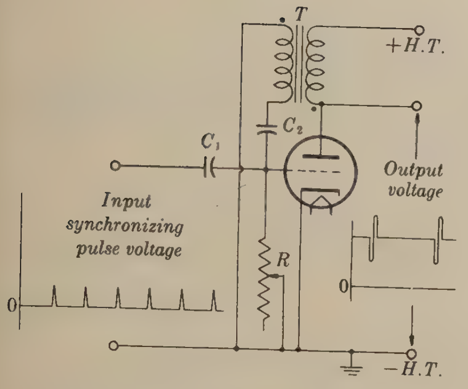


FIG.15-39 Synchronized blocking oscillator.

repeated. The time AB of the one cycle of sinusoidal oscillation is determined primarily by the transformer windings and their distributed capacitances. By using a well designed pulse transformer the duration of the complete oscillation at the anode may be made as small as a few microseconds. The period BC between the sinusoidal pulses and hence the number of such pulses per second is determined by the RC time constant. The output pulse may be coupled from the anode by means of a capacitor or by a third winding on the pulse transformer. This oscillator is called a *free-running blocking oscillator* because the duration of the quiescent period BC is determined by the circuit itself. The time AC is the *free-running period* of the oscillator.

2. SYNCHRONIZED BLOCKING OSCILLATOR. A free-running blocking oscillator may be synchronized to produce a given number of sinusoidal pulses per second by applying a series of positive synchronizing or trigger pulses to its grid. The circuit of Fig. 15-39 is adjusted to produce one output pulse for every five input pulses and so may be used as a *frequency divider*. The operation of the circuit is similar to that of the free-running blocking oscillator previously described. The variable grid resistor R is used to vary the free-running period. It is set so that the grid of the oscillator is beyond cut-off somewhat longer than five times the period of the input pulses, Fig. 15-40 (a). One of these pulses has initiated the blocking oscillator cycle of the circuit. This is indicated by pulse number 0 of graph (b) in Fig. 15-40. The grid is driven far beyond cut-off by the feedback through T and returns toward cut-off exponentially as shown in graph (c). At the same time the positive sync. pulses are applied to the grid of the valve through capacitor C_1. The actual grid voltage rise is the result of these two voltages and is shown in graph (c). Pulses 1, 2, 3 and 4 raise the grid potential momentarily but are not of sufficient amplitude to drive the grid above cut-off and initiate a new cycle. Pulse 5 raises the grid potential above cut-off and current begins to flow through the valve. The blocking oscillator cycle is therefore started, one output pulse being produced at the anode for every five input pulses applied to the grid.

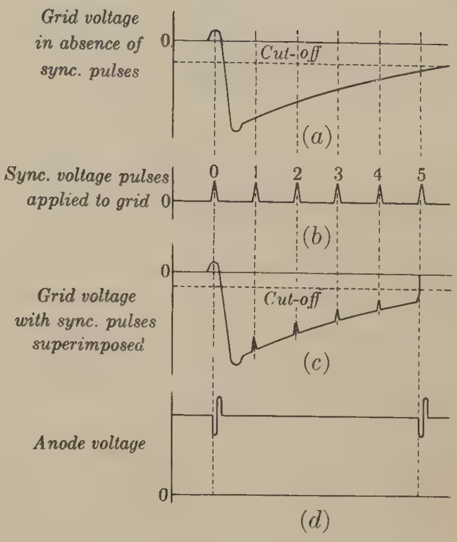


FIG.15-40 Operation of a synchronized blocking oscillator.

By decreasing the resistance of R, the time constant of the grid circuit is reduced and the grid voltage curve rises more steeply toward cut-off. In this way

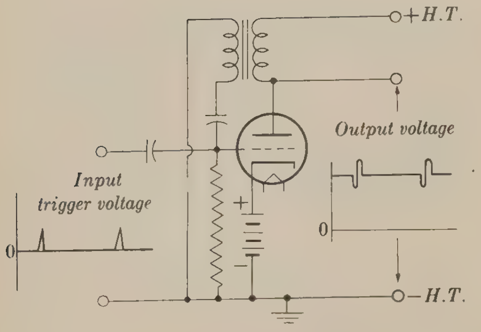


FIG. 15-41 Biased-off blocking oscillator with input and output waveforms.

every fourth, third, etc. sync. pulse applied to the grid may be made to produce one cycle of oscillation. For reliable operation of this circuit, the sync. pulse must be of uniform amplitude and regularly spaced. Negative sync. pulses applied to the anode may be used instead of positive sync. pulses to the grid. The coupling transformer reverses the polarity of the anode pulses and applies them to the grid as positive pulses.

3. BIASED-OFF BLOCKING OSCILLATOR.

Both the free-running type and the synchronized type of blocking oscillator produce oscillations at regular intervals with no input applied. A third type of oscillator shown in Fig. 15-41 produces a cycle of oscillation only when an input pulse is applied and remains non-conducting at other times, Figs. 15-42 (a). The valve is maintained in a biased-off condition by a large positive potential on the cathode or by a similar negative potential on the grid. A positive input pulse must be applied to the grid of sufficient amplitude to overcome this bias and the circuit executes one cycle of oscillation. At the end of the cycle, the bias maintains the grid beyond cut-off until another input pulse is applied.

By suitably adjusting the bias, this circuit may be used to provide sharp pulses at the anode from a sine wave input to the grid (Fig. 15-42). The time constant of the grid circuit should be arranged to provide extra bias to keep the valve well below cut-off from the time the blocking oscillator cycle is produced until the desired input pulse returns the grid above cut-off again. If this time constant is too short, a second blocking oscillator cycle may be initiated by an input pulse before the desired one. For example, in Fig. 15-42 each third input cycle initiates the oscillator cycle.

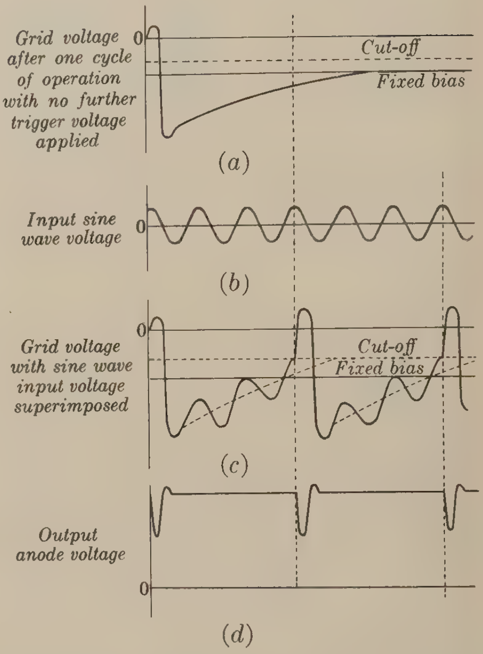

Grid voltage after one cycle of operation with no further trigger voltage applied
0
Cut-off
Fixed bias
(a)
Input sine wave voltage
0
(b)
Grid voltage with sine wave input voltage superimposed
0
Cut-off
Fixed bias
(c)
Output anode voltage
0
(d)

FIG. 15-42 Operation of a biased-off blocking oscillator with sine wave input.

E. *Thyratron Pulse Former*

A circuit which produces a very sharp positive pulse of considerable amplitude is shown in Fig. 15-43. A thyratron valve has a load resistor R_4 of small resistance connected in its cathode circuit and a large resistance R_3 connected in its anode circuit. The valve is in a normally biased-off condition due to the negative potential supplied to its grid by the bias battery. A storage capacitor C_2 is con-

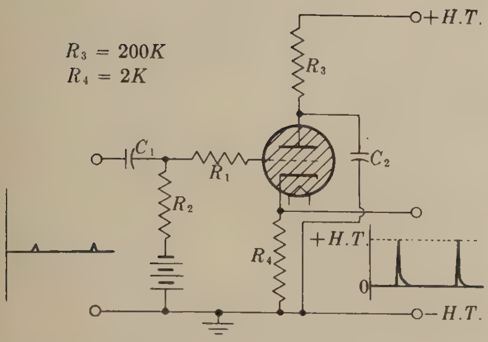


FIG.15-43 Thyratron pulse former.

nected across the valve and its load resistor.

When the H.T. is connected C_2 charges exponentially through R_3 to the full value of the supply voltage. A positive input pulse is applied to the grid of the valve of sufficient amplitude to overcome the grid bias. The gas within the thyratron ionizes and the valve conducts heavily. The resistance between cathode and anode falls to a very low value. The storage capacitor C_2 is now effectively connected across the cathode load resistor R_4 in series with the low resistance of the valve. C_2 quickly discharges through R_4 and the valve and most of the initial voltage on the capacitor is developed across R_4. As the time constant of this circuit is very small, the output pulse developed across R_4 is of very short duration but of very large amplitude.

As C_2 discharges, the anode potential falls. When the de-ionization potential is reached, the valve stops conducting and C_2 again begins to charge through R_3 to the supply voltage. The time constant of C_2 and R_3 is made sufficiently long that ionization cannot occur a second time before the input pulse is removed from the grid of the thyratron. On the other hand, R_3 must not be so large as to prevent C_2 from charging completely before a second firing pulse is applied to the grid. R_1 is a resistor which limits the flow of grid current in the thyratron when large input pulses are applied.

F. *Discharge Line*

A section of transmission line ordinarily used for conducting radio-frequency currents may be used to form a pulse of electrical energy of rectangular wave shape and extremely short duration. The line may be of the parallel conductor type or of the concentric line type as described in Chapter X. In practice however, an *artificial line* is used consisting of series inductors and shunt capacitors. In this way, the characteristic impedance may be made somewhat higher than that provided by most transmission lines and the physical length is very much reduced.

The operation of a discharge line is most readily understood by considering the effect of connecting a long section of concentric line to a battery as in Fig. 15-44. The battery has an e.m.f. of 200 volts and an internal resistance of 500 ohms. The line has a characteristic impedance of 500 ohms and matches the internal resistance of the battery. Before the switch is closed in the circuit the line is discharged. The inner and outer conductors are at the same potential. At the instant the switch is closed, electrons begin to flow from the negative terminal of the battery into the left hand end of the outer conductor. Electrons also flow from the central conductor back to the positive terminal of the battery. This

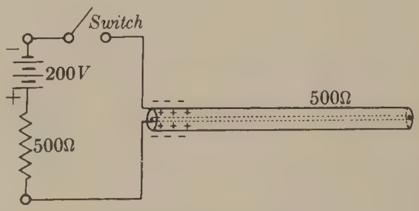


FIG.15-44 Discharge line charging circuit.

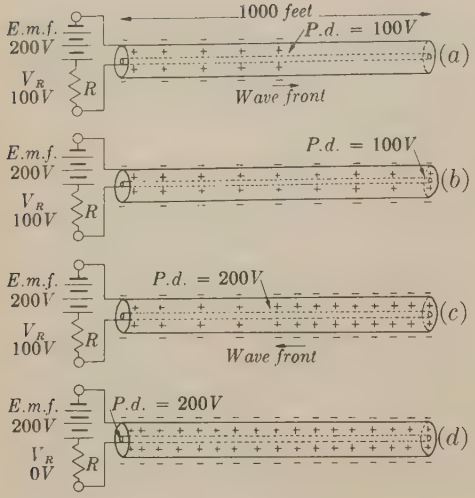


Fig.15-45 Charging process of a discharge line.

constitutes a flow between the inner and outer conductors. The series inductance in each element of length of the line maintains this current uniform at a value given by Ohm's Law, $I = \frac{V}{R} = \frac{200}{1000}$ amperes or $200mA$. The wavefront of current advances from the input end to the open end of the line at the speed of light which is approximately 1000 feet per microsecond.

The distribution of voltage must also be considered. Since the impedance of the line matches the internal impedance of the source, the e.m.f. of the battery divides equally between them. When the switch is closed, the terminal p.d. of the battery falls to 100 volts and a p.d. of 100 volts is built up between the inner and outer conductor of each successive element of the line. This wavefront of voltage advances down the line in phase with the current wavefront. The diagrams of Fig. 15-45 (a) and (b) show the distribution of charge, the flow of electrons and the voltages which exist in the circuit at the end of ½ microsecond and 1 microsecond respectively, along a line of length 1000 feet.

When the wavefront reaches the open end of the line, a current wavefront begins to move in the reverse direction back toward the battery. This current must be assumed to exist in order that the outgoing wave of current at the open end be cancelled because the net current flow at that point must be zero. Thus a new current wavefront moves from the right hand end of the line to the battery end in the next microsecond. Since the p.d. builds up as the current wavefront advances, the p.d. at the open end of the line must rise suddenly from zero to 100 volts to 200 volts. The potential of each successive element of the line rises from 100 volts to 200 volts until the wavefront reaches the battery, Fig. 15-45 (c). At this instant the current stops flowing, the p.d. across the internal resistance disappears and the 200-volt source is opposed by the 200 volts of the charged line, Fig. 15-45 (d).

The discharge of the line is a similar phenomenon but takes place in the reverse direction. Fig. 15-46 shows a transmission line 1000 feet long having a characteristic impedance of 500 ohms charged to a p.d. of 200 volts and connected through a switch to a resistor of 500 ohms. At the instant the switch is closed electrons begin to flow from the left end of the outer conductor through R to the inner conductor. The p.d. across the left end element falls from 200 volts to 100 volts and a p.d. of 100 volts is developed across the resistor. The discharge wave travels down the line to the right hand end in 1 microsecond discharging each element of the line to 100 volts and maintaining the p.d. across R at 100 volts. At the open end of the

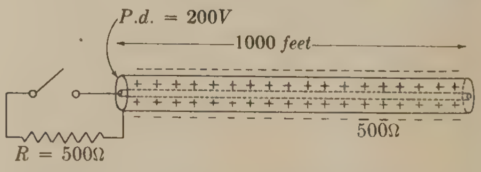


Fig.15-46 Discharge line discharging circuit.

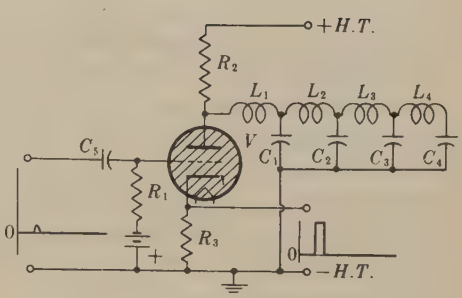


FIG. 15-47 Discharge line pulse-forming circuit.

line, a discharge wavefront then starts to move back toward the left end of the line. This discharges the line from 100 volts to zero in the next microsecond still maintaining the p.d. across R at 100 volts. A rectangular wave of voltage is therefore developed across R with an amplitude of 100 volts and a duration of 2 microseconds following the closing of the switch.

It must be emphasized that the phenomenon takes place as described only when the load resistance matches the characteristic impedance of the line. If these are mismatched the line is not completely discharged in a single discharge cycle. A current wavefront moves down the line and back again causing a partial discharge only. A redistribution of voltage then takes place between the line and the load and another discharge cycle occurs. This process is repeated and the line voltage falls in steps until it reaches zero.

The practical application of the discharge line as a pulse former is shown in Fig. 15-47. A simulated discharge line consisting of series inductors L_1 to L_4 and shunt capacitors C_1 to C_4 is connected between ground and the anode of a thyratron valve V. A load resistor R_3 is placed in the cathode circuit of the valve. The resitance of R_3 together with the resistance of the valve when conducting is made equal to the characteristic impedance of the discharge line.

The valve is kept in a non-conducting state by the bias battery in the grid circuit until a signal is applied to the grid. The capacitors in the discharge line become charged through R_2 to the value of the supply voltage. Since there is no current through the valve, the output voltage across R_3 is zero. A positive trigger pulse voltage is applied to the grid at the instant when the output pulse is to be developed. The grid bias is overcome causing the valve to conduct heavily. The p.d. to which the discharge line capacitors are charged falls to half its original value. An equal voltage is applied across R_3 in series with the valve. Since the resistance of the valve when conducting is very small, most of this p.d. is developed across R_3. The capacitors discharge successively from left to right to half their voltage, then from right to left to zero potential. The flow of current from the lower plates of these capacitors through R_3 and the thyratron to the upper plates of the capacitors maintains the p.d. across R_3. The output therefore is a positive pulse of rectangular waveform. The duration of this pulse is determined by the equivalent length of the discharge line and may be increased by adding additional sections. A pulse of this type may be made as short as one microsecond, or even less.

The amplitude of the pulse is determined by the value of the supply voltage. Discharge line circuits may be used at voltages of a few hundred volts, followed by pulse amplifiers to increase the power. Alternatively, a high-voltage line constructed with well insulated capacitors may be used to develop a pulse of several thousand volts amplitude. The output impedance is usually sufficiently low that a step-up transformer may be used to provide a further boost in voltage.

15.4 Square-wave generators. The circuits which are described in the following sections have been variously named for the functions they perform or their mode of operation. These names include multivibrators, relaxation oscillators, flip-flops, delay circuits, triggered circuits, relays, transitrons, etc., A slight modification in one of these circuits is frequently all that is necessary to change it to another type. For this reason, the circuits are not classified either as to purpose or operation. They have in common one characteristic namely that all produce waveforms of voltage which rise or fall very abruptly. For this reason they are frequently called *square-wave generators.*

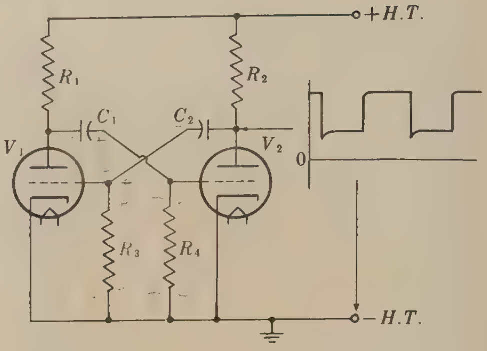


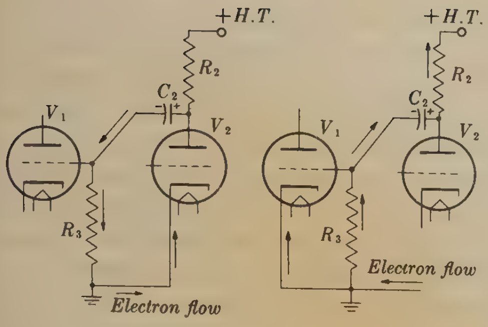
FIG.15-48 Symmetrical multivibrator.

A. Symmetrical Multivibrator

The basic multivibrator is shown in Fig. 15-48. It consists of two exactly similar triode amplifiers with the output of each resistance-capacitance coupled to the input of the other. R_1 and R_2 are the anode load resistors of the two valves. R_3 and R_4 are the grid-leak resistors and C_1 and C_2 the coupling capacitors. During normal operation of the circuit the two valves conduct current alternately. When one is conducting the other is cut off. A detailed explanation of the reasons for this behaviour and the factors which determine the length of each cut-off period follows.

When the heater and H.T. circuits are completed, current begins to flow through both valves. However, both valves will not continue to conduct exactly the same amount of current at every instant. Any slight unbalance in the circuit causes one valve to conduct more heavily than the other. Suppose a slight increase in current takes place through V_2. Then the p.d. across R_2 increases and the anode potential falls. C_2 therefore begins to discharge down through R_3 to the ground. This develops a p.d. across R_3 which applies a negative potential to the grid of V_1. The current through V_1 therefore decreases and the p.d. across R_1 decreases. The anode potential of V_1 rises, causing C_1 to charge up through R_4. This causes a p.d. across R_4 which applies a positive potential to the grid of V_2. The current through V_2 which was assumed to increase slightly at the first instant, is now made to increase greatly due to the amplified positive potential on its grid. The process is cumulative with the result that the grid potential of V_1 is driven far beyond cut-off while that of V_2 is made somewhat positive. Grid current flows in V_2 to charge C_1 so that the positive potential of the grid of V_2 is limited to a smaller value than the corresponding negative potential on the grid of V_1. The voltage changes at the grids and anodes of the two valves which are described as if they occurred slowly, actually take place extremely rapidly due to the amplification of the valves. The switching action may be completed in a matter of microseconds in a well designed multivibrator. The anode potential of V_1 rises almost to H.T. and becomes steady because the valve is cut off. The anode potential of V_2 is low because the valve is conducting heavily and there is a large p.d. across R_2 and is steady since the grid has become steady at a small positive potential.

The circuit now undergoes a period of relaxation during which the grid potentials tend to return to zero. C_2 which at



(*a*) (*b*)

Fig. 15-49 Electron flow through the coupling circuit of a multivibrator (a) Coupling capacitor discharging (b) Coupling capacitor charging.

first began to discharge rapidly through R_3 gradually discharges more slowly since the anode potential of V_2 is at a low steady value. The p.d. across R_3 decreases exponentially at a rate determined by the time constant of C_2R_3 and the resistance of V_2 when conducting. Thus the negative grid potential of V_1 rises toward cut-off. When this point is reached current begins to flow through R_4 and V_1 making the grid potential of V_2 negative. The anode current of V_2 is reduced and the anode potential rises. C_2 begins to charge through R_3 making the grid of V_1 positive. The switching action is similar to that previously described for V_2. In this case V_2 is cut off and V_1 is made to conduct heavily. The anode potential of V_2 rises to H.T. and that of V_1 is low. A second relaxation period then begins during which the grid potential of V_2 returns to cut-off exponentially. The duration of this period depends upon the time constant of C_1 and R_4. When the grid potential of V_2 reaches cut-off, current again begins to flow through this valve. This initiates a new switching action which completes the multivibrator cycle.

The charge and discharge paths of one of the coupling capacitors is shown in Fig. 15-49 (a) and (b). The diagram (a) indicates the direction of electron flow which discharges C_2 during the relaxation period of the first half cycle. Diagram (b) shows the path of electrons which charge C_2 during the second half of the cycle. Since grid current is flowing in V_1, C_2 charges quickly through the low impedance of the grid-cathode circuit.

A circuit of this type is called a *free-running multivibrator* because its frequency is determined by the values of the components which make up its own circuit. That is, it operates continuously without the aid of trigger pulses.

The duration of each relaxation period is determined by the time required for the grid to return to cut-off. In this multivibrator the duration of one relaxation period is the same as the other and the multivibrator is called a *symmetrical multivibrator*. This duration is usually about one time constant of the coupling capacitor and grid-leak resistor. The period of one complete oscillation is therefore given by the formula

$$T = C_1R_4 + C_2R_3$$

and the frequency by the formula

$$f = \frac{1}{C_1R_4 + C_2R_3}.$$

The frequency of oscillation of the multivibrator may be increased by reducing the resistance of either or both of R_3 and R_4, or by reducing the capacitance of either or both of C_1 and C_2.

The grid and the anode voltage changes are shown graphically in Fig. 15-50. The region *ab* in graph (a) indicates the rapid fall of grid potential which cuts off the current through V_1. At the same time its anode potential rises toward H.T. as shown by the region *ab* in graph (b). In graphs (c) and (d) the grid potential of V_2 is shown slightly positive while the anode potential falls abruptly due to the heavy current flowing through this valve. The relaxation period is indicated by the region *bc* in graph (a) during which the

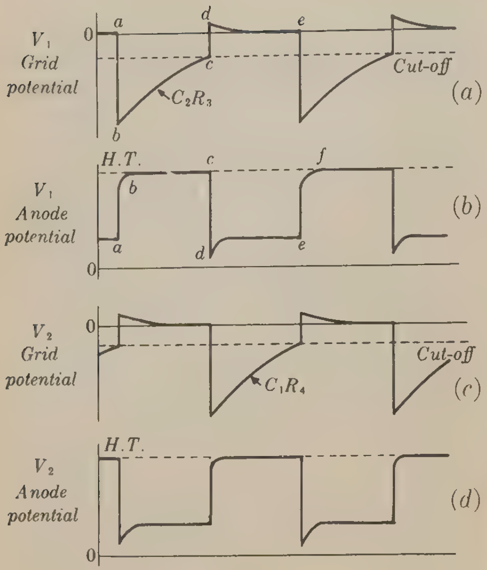
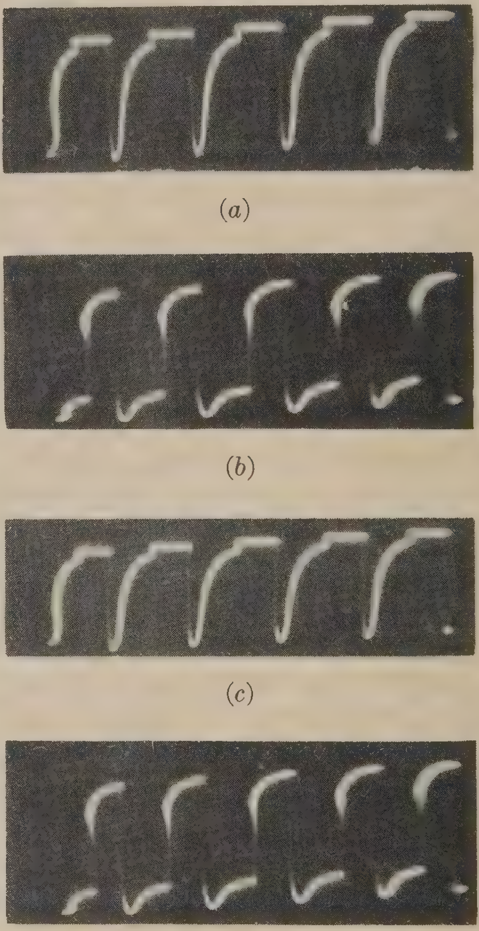


FIG.15-50 Operation of a symmetrical multivibrator.

grid potential of V_1 returns to cut-off. The anode potential of V_1 remains at H.T. during this period. The anode potential of V_2 rises only slightly as its grid voltage drops to zero. The second switching action takes place from *c* to *d* in graph (a) as the grid of V_1 is driven positive. The anode potential of this valve falls steeply from *c* to *d*. The grid of V_2 is driven negative beyond cut-off and its anode potential rises to H.T. The second relaxation period is completed in the time indicated by *de* in graph (a). During this period the grid potential of V_2 rises to its cut-off value to complete the cycle.

The anode waveforms depart considerably from rectangular shapes because of two factors. In the region *ab* of graph (b) the valve may be cut off instantaneously but the anode potential cannot rise more rapidly than the anode-grid coupling capacitor can charge up. This causes a certain degree of integration of the square waveform which is indicated by the exponential rise of *ab* in graph (b).

On the other half of the cycle, the anode potential rises slightly from *d* to *e*, graph (b), because the grid potential decreases from a positive value at *d* to zero at *e*, graph (a). In order to maintain an abrupt switching action the circuit must be capable of amplification over the wide range of harmonic frequencies contained in the square waveform. This requires that the circuit be designed as a video amplifier. Multivibrators of this type may be made to operate at frequencies as low as one cycle per minute and as high as 100,000 cycles per second.



(*a*)

(*b*)

(*c*)

(*d*)

FIG.15-51 Oscillograms of voltage waveforms in a symmetrical multivibrator.

Fig. 15-51 shows actual photographs of the multivibrator waveforms as they appear on an oscilloscope.

B. Unsymmetrical Multivibrator

An *unsymmetrical* or *unbalanced multivibrator* is one in which the coupling capacitor and grid-leak resistor associated with one valve has a longer time constant than that associated with the other valve. The circuit diagram of such a multivibrator is the same as that of the symmetrical multivibrator shown in Fig. 15-48. If the time constant of C_1R_4 is greater than that of C_2R_3, the time required for the grid potential of V_2 to return to cut-off will be greater than the corresponding time for V_1. Hence the anode potential of V_1 is high for a comparatively short time and low for a longer time during each cycle. The reverse is true at the anode of V_2 as shown in Fig. 15-52. A circuit of this type may be used to produce a square pulse of voltage of any required duration less than the period of the whole multivibrator cycle which can have any required repetition rate.

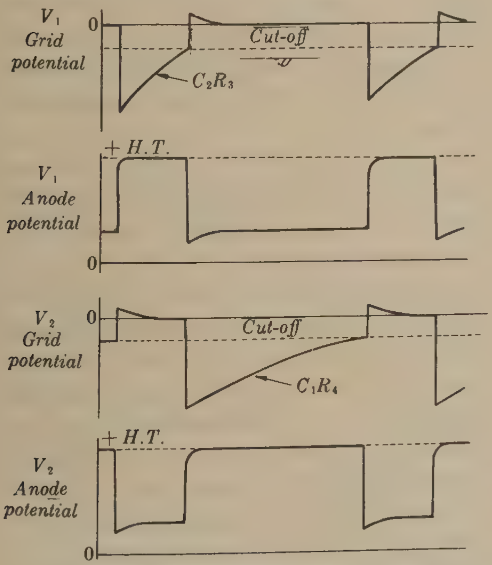


FIG. 15-52 Operation of an unsymmetrical multivibrator.

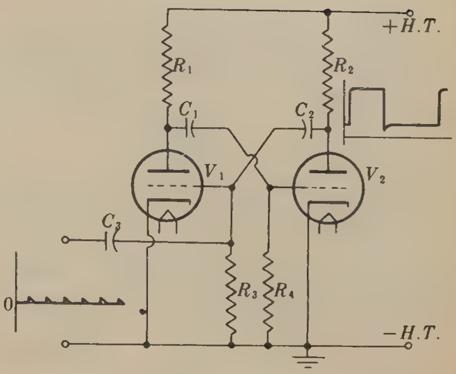


FIG. 15-53 Synchronized multivibrator.

C. Synchronized Multivibrator

The frequency stability of free-running multivibrators is poor and consequently they are not widely used as such in radar equipment. Multivibrators are usually made to operate at a specific frequency by a synchronizing voltage applied to one of the electrodes. Almost any waveform of sync. voltage may be used. A sine wave or a series of pulses is usually most convenient. An unsymmetrical multivibrator is shown in Fig. 15-53 in which positive sync. pulses of uniform amplitude and frequency are applied to the grid of V_1 through capacitor C_3. The multivibrator is made to begin a new cycle of operations by every fifth sync. pulse applied to it. That is, the multivibrator operates at a frequency equal to one fifth of the repetition rate of the sync. pulses.

Voltage waveforms for this circuit are shown in Fig. 17-54. Graph (a) shows the waveform of voltage at the grid of V_1 when no synchronizing pulses are applied. The duration of the period *ab* in graph (a) is determined by the time constant of C_1R_4 while the time *bc* is controlled by the time constant of C_2R_3. The sync. pulses applied to the grid of V_1 are shown in graph (b). These pulses are superimposed upon the grid voltage as shown in graph (c). The first sync. pulse applied

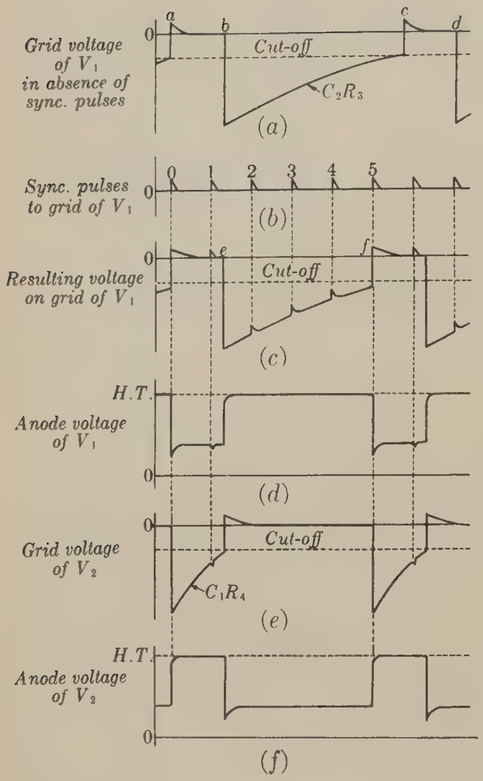


FIG.15-54 Operation of a synchronized multivibrator.

after the beginning of the cycle makes the grid of V_1 slightly positive and its anode slightly less positive for an instant. This pulse, coupled through C_1 to the grid of V_2 as a negative pulse has no effect on the duration of this half cycle as the grid potential of V_2 is already below cut-off. Similarly pulses 2, 3 and 4 in graph (c) have no effect on the duration of the other half cycle. They raise the grid potential momentarily as indicated in graph (c) but are not of sufficient amplitude to drive the grid above the cut-off value. Pulse 5 however, raises the grid potential of V_1 above cut-off. Immediately current begins to flow through V_1, its anode potential begins to fall and the grid of V_2 is driven negative. The switching action initiated by the fifth pulse is carried on by the multivibrator circuit and a new cycle is started. The effect of the sync. pulses then is to reduce the duration of one half cycle from that indicated by *bc* in graph (a) to that indicated by *ef* in graph (c). Graphs (d), (e) and (f) have been included to show that all the sync. pulses except every fifth have no effect on the duration of the cycle. Pulses 2, 3 and 4 do not appear at the anode of V_1. They do not cause anode current to flow through the valve and so the anode potential remains at H.T.

The multivibrator may be locked in with every fourth sync. pulse by reducing the resistance of R_3. This decreases the time constant of C_2R_3 and causes the grid potential of V_1 to rise toward cut-off more steeply. The fourth sync. pulse may therefore become effective in driving the grid above the cut-off potential and starting a new cycle. Circuits of this type may be used as frequency dividers with a division ratio as high as 1:10. The sync. pulses must be of uniform amplitude and repetition rate if the multivibrator is to be accurately controlled.

D. *Cathode-Coupled Multivibrator*

The *cathode-coupled multivibrator* shown in Fig. 15-55 differs from the types previously described in that the anode voltage of V_2 is not coupled to the grid of V_1. Instead, a switching action is provided by coupling the cathodes together through the common resistor R_3. The grid of V_1 is maintained at ground potential by R_4. The duration of one half cycle is determined by the discharge time of C_1 through R_5, R_3 and V_1. The duration of the other half cycle is determined by the charging time of C_1 through R_3, R_1 and the grid-cathode circuit of V_2 as well as through R_1 and R_5. A detailed description of the operation follows, together with voltage waveforms in Fig. 15-56.

If the L.T. and H.T. supplies are connected but the switch S is open, capacitor

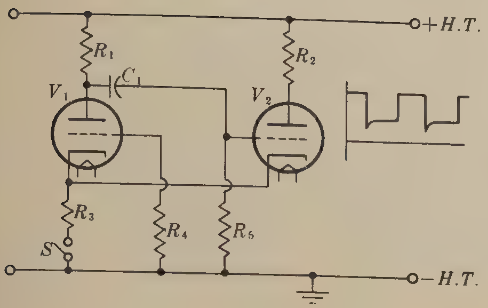


FIG. 15-55 Cathode-coupled multivibrator.

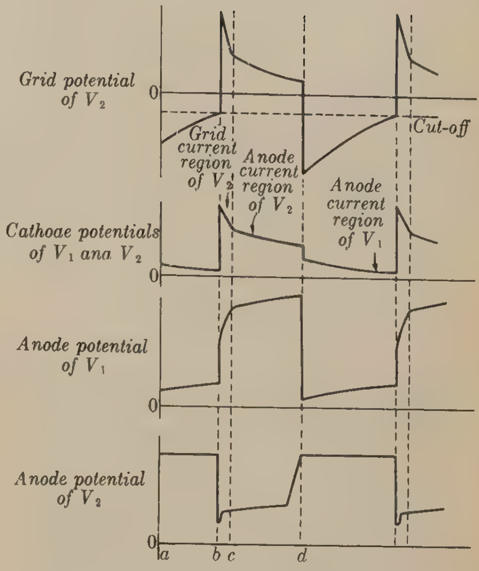


FIG. 15-56 Operation of a cathode-coupled multivibrator.

C_1 will charge up to the full value of the supply voltage through R_1 and R_5. The anode of V_1 will be at H.T. potential because no current will flow through R_1 after C_1 is charged. The anode of V_2 is at H.T. potential since the cathode circuit is broken. Both grids will be at zero potential due to R_4 and R_5 which are returned to ground.

When S is closed, current may begin to flow through both valves developing voltages across R_1 and R_2. This causes the anode potentials of V_1 and V_2 to fall from their initial H.T. potential. A fall in anode potential of V_1 causes C_1 to begin discharging through R_5 producing a negative voltage on the grid of V_2. The current through this valve decreases and so the p.d. across the common cathode resistor R_3 also decreases. A less positive potential on the cathode of V_1 is equivalent to a more positive potential on the grid. V_1 conducts more heavily. Its anode potential drops sharply, and C_1 applies a more negative potential to the grid of V_2. This process is cumulative and the current through V_2 is quickly cut off. Although the sequence of events is described as though it takes place gradually, the switching action is completed in a very few microseconds as indicated by the steep waveforms of Fig. 15-56.

A relaxation period now takes place with V_1 conducting and V_2 cut off. The grid potential of V_2 rises exponentially as C_1 discharges through R_5,R_3 and V_1 since the anode potential of V_1 is at a low steady value. When cut-off potential is reached, current again begins to flow through V_2. A second switching action now takes place. The increased current through R_3 causes the cathode potential to rise. This is equivalent to a negative potential on the grid of V_1 and results in a decrease in current through this valve. Its anode potential therefore rises and C_1 begins to charge through R_5. This places a positive potential on the grid of V_2 causing it to conduct more heavily. When the grid of V_2 is more positive than its cathode grid current flows and C_1 charges more quickly through R_3 and the grid-cathode circuit of V_2 and R_1. The cathode potential rises sufficiently to cut off the current through V_1 and V_2 is left in a heavily conducting state. This is indicated by time b in Fig. 15-56. The relaxation period of this half cycle consists of two parts. During the time indicated by bc, the capacitor C_1 charges by drawing electrons from the grid of V_2. The resistance of R_3, V_2 and R_1 is

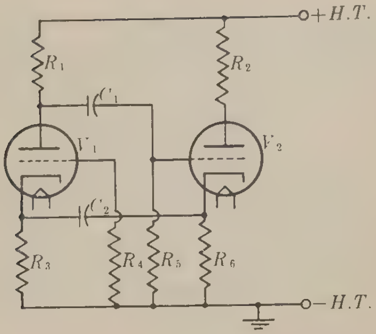


FIG. 15-57 Modified cathode-coupled multivibrator.

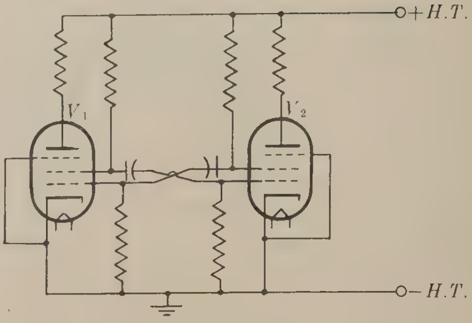


FIG. 15-58 Electron-coupled multivibrator.

comparatively small so that the positive grid potential decreases rapidly. At *c*, the positive grid potential equals the positive cathode potential and grid current ceases to flow. C_1 continues to charge through the greater resistance of R_5 instead of through R_3 and the cathode-to-grid circuit of the valve V_2. This is indicated by the more gradual fall of grid potential from *c* to *d*. As the grid voltage decreases, the current through V_2 decreases and the cathode potential becomes less positive. At *d*, the cathode potential has fallen to the point where current again begins to flow through V_1 and a new cycle is started.

A modification of the cathode-coupled multivibrator is shown in Fig. 15-57 in which separate cathode resistors R_3 and R_6 are employed for V_1 and V_2, with capacitive coupling between them via C_2. The operation of the circuit is similar to that of the previous circuit during the time V_2 is cut off. During this time there is no voltage across R_6 and C_2 is charged to the voltage across R_3, the positive plate being connected to the cathode of V_1. The duration of this half cycle is determined as before by the discharge time of C_1 through R_5, R_3 and V_1. When the grid potential of V_2 rises above cut-off, current begins to flow through R_6 and the valve causing a rise in cathode potential. This causes C_2 to begin discharging, the electron flow being up through R_3, V_2 and R_2. The rise in the cathode voltage of V_1 due to the extra electron flow through R_3 causes the valve V_1 to conduct less current and its anode potential rises. C_1 then applies a positive voltage to the grid of V_2. The regeneration action causes V_2 to conduct heavily while V_1 is cut off. Meanwhile C_2 has completed its discharge and is charging up through R_3, V_2 and R_2. The voltage across R_3 produced by this charging current maintains V_1 cut off. The duration of this relaxation period is determined by the charging time of C_2 through R_3, V_2 and R_2. As C_2 charges, the p.d. across R_3 decreases so that eventually current begins to flow through V_1 again.

E. Electron-Coupled Multivibrator

The *electron-coupled multivibrator* shown in Fig. 15-58 employs pentode valves instead of triodes. The cathode, control grid and screen grid of each valve perform the functions of cathode, grid and anode of the triode circuits in the basic multivibrator. When the control grid of the conducting valve is positive permitting a large flow of current to the screen, the anode current is also a maximum. This develops a large p.d. across the anode load resistor and the anode potential is low. During the

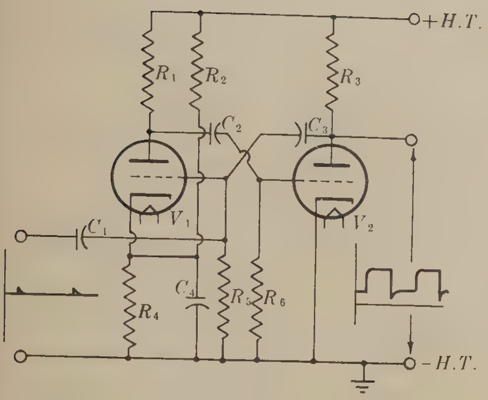


FIG. 15-59 Biased-off multivibrator.

other half of the cycle when the control grid potential is below cut-off, no current flows to either the screen or the anode. The anode potential therefore rises to the supply voltage. The waveform produced at the anode is similar to that produced at the screen grid. The suppressor grid, which remains at ground potential, shields the multivibrator section of the valve from the anode circuit. In this way a square wave of voltage may be supplied by either anode to the following circuit without loading the multivibrator. Thus the frequency of oscillation is reasonably independent of the output circuit.

F. Biased-Off Multivibrator

A *biased-off multivibrator* differs from a free-running multivibrator in that it remains in a steady state until triggered by a pulse applied to one of its electrodes. When such a pulse is received, the multivibrator goes through one cycle of operation and then remains in a quiescent state until triggered again. For this reason it is sometimes called a *one-shot multivibrator*.

Fig. 15-59 shows a multivibrator of this type. V_1 is biased so that no current flows through this valve during its steady state condition. The grid is returned to ground potential through R_5 while the cathode is maintained at a positive potential from the voltage divider R_2, R_4 across the H.T. supply. This grid-cathode difference in potential is made to exceed the cut-off voltage of the valve so that no current flows through V_1. Its anode potential equals the H.T. voltage since no p.d. exists across R_1 under these conditions.

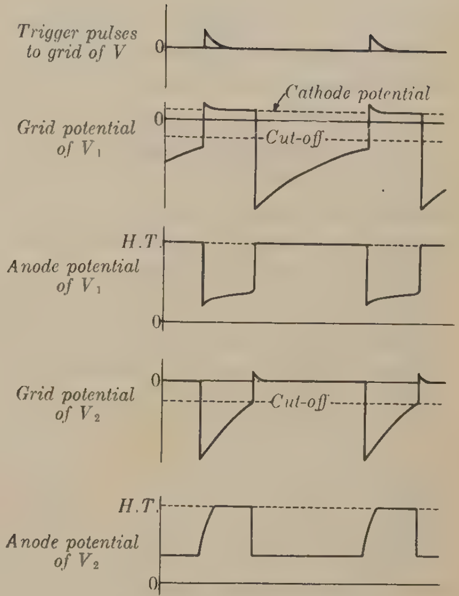


FIG. 15-60 Operation of biased-off multivibrator.

V_2 is made to conduct heavily by returning both its grid and cathode to ground potential. The flow of anode current through R_3 develops a large p.d. across this resistor so that the anode potential is relatively low. A stable or equilibrium condition therefore exists with V_2 conducting and V_1 not conducting current.

A positive trigger pulse is applied via C_1 to the grid of V_1 of sufficient amplitude to overcome the bias and allow the valve to start conducting (Fig. 15-60). The flow of current through R_1 develops a p.d. across this resistor causing the anode potential to fall. C_2 therefore begins to discharge through R_6 and applies a negative potential to the grid of V_2. The current flowing through this valve decreases and its anode potential rises. C_3 therefore begins to charge through R_5 and applies a positive potential to the grid of V_1. The current through this valve which was initiated by the sync. pulse, is now increased by the positive grid potential due to the multivibrator action. The action is cumulative and takes place very rapidly because of the amplification of both valves. V_1 is made to conduct very heavily and V_2 is cut off. The anode potential of V_1 falls to a minimum value while that of V_2 rises to the supply voltage.

A relaxation period follows during which C_3 quickly completes charging due to grid current flow in V_1. C_2 continues to discharge through R_6 so that the negative potential on the grid of V_2 decreases exponentially. When the grid potential reaches cut-off, current again begins to flow through V_2. Its anode potential falls and C_3 discharges through R_5, making the grid of V_1 negative. The current through this valve decreases and its anode potential rises. This causes C_2 to begin charging through R_6 and makes the grid potential of V_2 positive. The cumulative action takes place again, this time causing V_2 to conduct current and V_1 to be cut off. When this condition is reached, V_1 remains in a non-conducting state due to the positive bias applied to its cathode. The by-pass capacitor C_4 maintains the cathode of this valve at a-c ground potential during the cycle.

The output voltage of the multivibrator is a negative square pulse from the anode of V_1 or a positive square pulse from the anode of V_2. This square pulse begins when the trigger pulse is applied and continues for a period determined by the time constant of C_2 and R_6, as indicated by the graphs of Fig. 15-60.

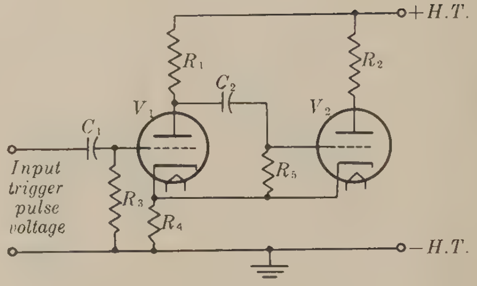


FIG. 15-61 Cathode-coupled, biased-off multivibrator.

Biased-off multivibrators may vary considerably as to the method of providing the bias and coupling. A biased-off multivibrator with cathode coupling is shown in Fig. 15-61. V_2 is the conducting valve as its grid is returned to the cathode for zero bias. The cathode to anode current of this valve flows through the common cathode resistor making both cathodes positive to ground. The grid of V_1 is returned to ground through the grid-leak resistor. The p.d. across the cathode resistor provides the bias which maintains V_1 in a non-conducting state until a positive trigger pulse is applied. The multivibrator action is similar to that described for the cathode-coupled and biased-off multivibrators. The duration of the cut-off period of V_2 is determined by the discharge time of C_2 through R_5 and V_1. When V_2 conducts V_1 is driven well beyond cut-off. The time required for V_1 to return to its original biased-off state is determined by the charging time of C_2 through R_4, R_5 and V_2, and R_1. Beyond this recovery period, the trigger pulse becomes effective in starting a new cycle. Thus the recovery period must be less than the period between trigger pulses.

G. Eccles-Jordan Circuit

The *Eccles-Jordan circuit* shown in Fig. 15-62 has two conditions of stable operation in which one valve conducts heavily and the other is cut off. When a

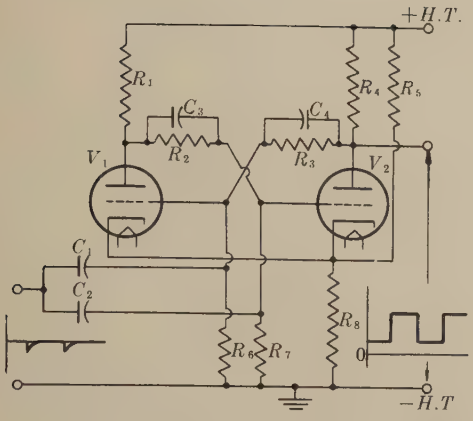


FIG. 15-62 Eccles-Jordan circuit.

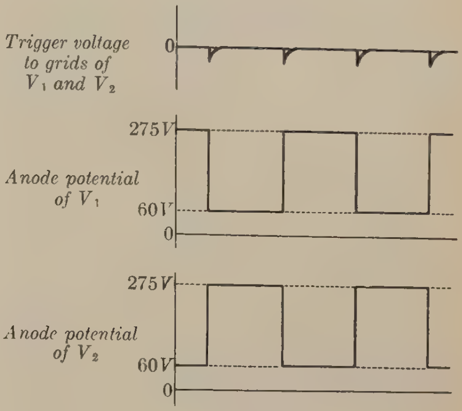


FIG. 15-63 Operation of Eccles-Jordan circuit.

negative pulse is applied to the two grids, the conducting valve is cut off and the non-conducting valve is made to conduct. This condition continues until a second trigger pulse is applied causing the valves to reverse their operation again. The circuit resembles a symmetrical multivibrator with two important exceptions.

(i) Direct resistance coupling is provided between the anode of each valve and the grid of the other.

(ii) The common cathode potential is positive with respect to ground at all times by an amount greater than the cut-off voltage of the valves.

A detailed description of the operation of the Eccles-Jordan circuit shown follows.

V_1 is assumed to be the non-conducting or *off* valve and V_2 is the *on* valve. The cathode potential of both valves is 30 volts since R_8 is part of the voltage divider R_5,R_8 across the H.T. supply and since the anode current of the *on* valve flows through R_8. R_1, R_2 and R_7 constitute another voltage divider across the H.T. supply which sets the anode potential of V_1 at 275 volts and the grid voltage of V_2 at +31 volts. The flow of grid current in V_2 through R_2 and R_1 is partly responsible for setting these potentials, as the grid of V_2 is one volt positive to its cathode. The current flowing through V_2 develops a large p.d. across R_4. This current, together with that from the voltage divider R_6, R_3 and R_4, sets the anode potential of V_2 at 60 volts which is much lower than that of V_1. Consequently the grid potential of V_1 is also low, being about 15 volts. This grid potential is 15 volts negative to the cathode. This bias is greater than the cut-off voltage of the valve. V_1 remains biased off and V_2 continues to conduct current until a trigger pulse is received.

The trigger pulse is a negative voltage of short duration (Fig. 15-63) applied by way of the coupling capacitors C_1 and C_2 to the grids of V_1 and V_2 simultaneously. This pulse has no effect on V_1 which is already cut off. It drives the grid of V_2 negative, reducing the flow of current through this valve. The p.d. across R_4 is reduced and the anode potential rises. The rising voltage is fed back through C_4R_3 and R_6 to the grid of V_1, overcoming its bias and causing it to conduct current. Its anode potential therefore falls, applying a negative voltage to the grid of V_2 through C_3R_2 and R_7. Thus the current through V_2 which at first was reduced by the trigger pulse, is now further reduced by the regenerative action of the circuit. The amplification of both valves causes V_1 to conduct heavily and V_2 to be cut off. The circuit remains in this condition because of the reversal of the two anode voltages and consequently the reversal of the two grid voltages.

C_3 and C_4 are omitted in some circuits of this type. The sudden change in anode voltage divides in the ratio of the coupling resistors R_3, R_6 or R_2, R_7 in the absence of the capacitor. In a typical example about one quarter of the change in voltage at the anode is applied to the grid of the opposite valve. With the capacitor connected in the circuit, a larger fraction of the change in the anode voltage is applied to the opposite grid, because the p.d. across the capacitor cannot change instantaneously. Thus the coupling capacitors serve to sharpen the triggering action of the valves. The resistor combination provides d-c coupling to maintain steady conditions between pulses, while the capacitor provides a-c coupling during the triggering period.

A square wave of voltage is obtained at either anode. The anode voltages are 180° out of phase with each other, as indicated by the diagrams of Fig. 15-63.

H. Transitron

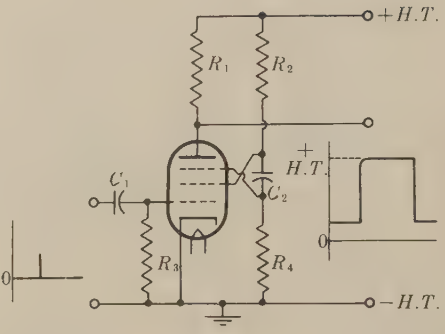


FIG. 15-64 Transitron square-wave generator.

A square-wave generator employing a single pentode valve is shown in Fig. 15-64. The circuit is known as a *transitron*. The suppressor of the valve, which is ordinarily connected to the cathode, is coupled through C_2 to the screen grid. It is this unusual circuit connection which makes regeneration possible. Since the suppressor is mounted between the anode and screen, it has dissimilar effects on the currents flowing to these two electrodes. A positive voltage on the suppressor causes it to attract electrons from the cathode through the screen to the anode. This increases the anode current. A negative voltage on the suppressor makes it repel electrons toward the screen. This causes an increase in screen current and a decrease in anode current without any appreciable change in the total current flowing through the valve.

The grid is operated at cathode potential or biased slightly negative. When no signal is applied, the cathode current through the valve divides, part flowing to the screen and part to the anode. When steady conditions have been reached, the anode and screen potentials remain at fixed positive values. The suppressor operates at ground potential due to R_4 and the capacitor C_2 is charged to a p.d. equal to the screen voltage.

When a sharp positive pulse is applied to the control grid the current through the valve increases. The increase in screen current causes the screen potential to fall. C_2 begins to discharge through R_4 and the valve, developing a negative potential at the upper end of R_4 which is applied to the suppressor. The suppressor repels electrons in the region between the screen and anode. This causes a further increase in screen current and a reduction of anode current. This cycle of events which was started by the trigger pulse on the control grid is now carried on by regeneration from the screen to the suppressor circuits. It takes place very quickly, causing the anode current to be cut off completely and the screen current to reach a maximum. During this time the anode voltage rises suddenly to H.T. Regeneration in this direction stops when all the cathode current is flowing to the screen grid.

A period of relaxation follows, during which the negative potential on the suppressor decreases as C_2 discharges through R_4 and the valve. When the suppressor potential becomes sufficiently less negative as the rate of discharge of C_2 decreases, current again begins to flow to the anode circuit. The screen current therefore decreases and the screen potential rises. C_2 now begins to charge through R_4 and R_2 making the suppressor positive. The anode current increases rapidly and the screen current decreases until both have resumed the values they had before the trigger pulse was applied. The output voltage, taken from the anode of the valve is a positive square wave which rises suddenly when the trigger pulse is applied and falls sharply at a later time. The duration of the output square pulse is determined by the discharge time of C_2 through R_4 and the valve and may be varied by adjusting R_4.

I. *Transitron Sine-Wave Oscillator*

A slight modification of the transitron square-wave generator just described is all that is required to make it generate a sine wave. This is made possible due to the fact that a square wave of voltage consists of a sine wave at a fundamental frequency and an infinite number of odd harmonics at the proper amplitudes. The fundamental frequency is the frequency of the square wave. The circuit modification causes just one of the sine wave components to be selected as the frequency of oscillation.

The transitron circuit of Fig. 15-65 contains a high-Q oscillatory circuit L_1C_1, instead of a resistor in the screen circuit. This has the effect of making regeneration very large at the resonant frequency to which this circuit is tuned. Voltages at all other frequencies are so small as to be negligible in the output, so that a sine wave of voltage is developed across L_1C_1. When the switch S is closed in the circuit, current flows to the anode and screen circuits. The increasing screen current develops a p.d. across L_1 which charges C_1. Thus energy is supplied to the oscillatory circuit and oscillations are started. As the screen potential falls, due to the increasing voltage across L_1, C_2 discharges somewhat through R_2 causing a negative voltage at the upper end of R_2. This negative potential is applied to the suppressor grid, causing the anode current to decrease and the screen current to increase. This supplies more energy to the oscillatory circuit. The effect is cumulative to the point of screen current saturation. The oscillation then reverses in L_1C_1 and the screen potential rises. This in turn causes C_2 to re-charge somewhat, producing a positive voltage at the top of R and hence on the suppressor. The screen current therefore decreases, a condition necessary to prevent degeneration in L_1C_1 during this half cycle. The suppressor therefore causes the screen current to increase and decrease at the frequency of the oscillations in L_1C_1. This is the same condition which maintains oscillation in the anode circuit of a series-feed oscillator. The frequency of oscillation is controlled by L_1C_1 and may be varied

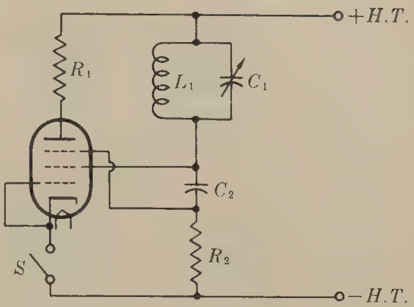


FIG.15-65 Transitron sine-wave oscillator.

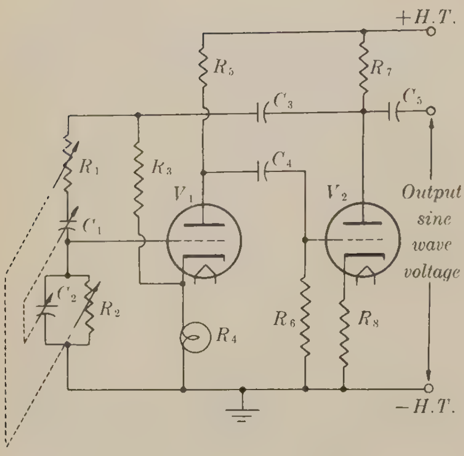


FIG.15-66 Wien-bridge oscillator.

by adjustment of either of these components.

15.5 Resistance-capacitance sine-wave oscillators. A two-valve amplifier circuit in which the output of each valve is resistance-capacitance coupled to the input of the other valve oscillates and generates a square wave of voltage. Since the square wave contains sinusoidal components at many frequencies regeneration must be taking place at all of these frequencies. This oscillator may be modified to produce a sine wave output by using a frequency-selective coupling network to eliminate the amount of regeneration required for oscillation at all frequencies except one. A *Wien-bridge* oscillator is an example of this type of circuit (Fig. 15-66). This oscillator is useful for producing an audio-frequency output voltage of almost constant amplitude and sinusoidal waveform over the entire range of audio frequencies.

V_1 is an amplifier coupled by C_4 to the grid of V_2. The output of amplifier V_2 is in turn coupled through C_3 to the input of V_1. This voltage is impressed upon the frequency-selective coupling network which is a Wien-bridge consisting of R_1, C_1, R_2C_2, R_3, and R_4. The voltage developed across the centre of the bridge is the feedback voltage applied between grid and cathode of V_1 as shown in Fig. 15-67. In the following paragraphs this feedback is considered to consist of two parts, a regenerative part and a degenerative part. The output of V_2 provides these two parts.

The degenerative or negative feedback is applied from the anode of V_2 to the cathode of V_1 through C_3, R_3 and R_4. When the grid potential of V_1 rises, its anode potential falls. This falling potential is applied to the grid of V_2 causing its anode potential to rise. The rising potential from V_2 is applied by means of the voltage divider R_3, R_4 to the cathode of V_1. But a rising cathode potential on V_1 is equivalent to a fall in its grid potential. Thus there is degenerative feedback which remains constant in amplitude and phase for all frequencies since the resistances of R_3 and R_4 are constant (Fig. 15-68).

Regenerative or positive feedback is

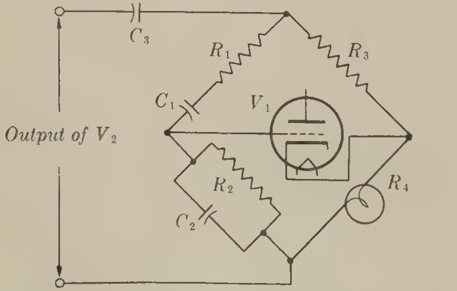


FIG.15-67 Frequency-selective bridge network in the Wien-bridge oscillator.

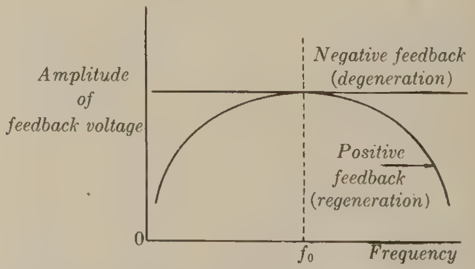


FIG.15-68 Graph of feedback voltage vs. frequency in the Wien-bridge oscillator.

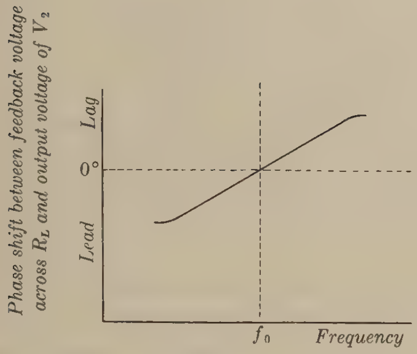


FIG. 15-69 Graph of phase shift between input to V_1 and output of V_2.

applied from the anode of V_2 to the grid of V_1 through C_1, R_1 and C_2R_2. As before, a rise in the grid potential of V_1 causes a rise in the anode potential of V_2. This in turn causes a rise in voltage across R_2 and C_2 and raises the grid potential of V_1 further. Thus there is regenerative feedback across R_2 which varies in amplitude and phase with varying frequency due to the presence of C_1 and C_2. The ability of this regenerative feedback to overcome the degenerative feedback and cause oscillation is greatest when its amplitude is a maximum and it is in phase with the output of V_2.

The manner in which the amplitude of the regenerative feedback varies with frequency is shown in Fig. 15-68. If the feedback voltage were of high frequency, the reactance of C_2 would be small compared with R_1 and R_2. Most of this voltage would appear across R_1 and only a small proportion would be applied to V_1. On the other hand, if the feedback voltage were of low frequency the reactance of C_1 would be large in comparison with the resistance of R_1 and R_2. Consequently most of the feedback voltage would be developed across C_1 and very little would be applied to V_1. At some intermediate frequency f_0 therefore, the positive feedback from V_2 to V_1 becomes a maximum. The curve is rather flat in the region about f_0.

It may be shown that at this same frequency f_0, the feedback voltage across R_2 is in phase with the output voltage from V_2 (Fig. 15-69).

The values of the bridge components are so chosen that the effect of the regenerative feedback having maximum amplitude combined with the in-phase condition cause the degenerative feedback to be just overcome so that oscillation takes place at the frequency f_o. The frequency at which the circuit oscillates is given by

$$f_0 = \frac{1}{2\pi\sqrt{R_1C_1R_2C_2}}$$

In the practical circuit R_1 and R_2 are made equal and C_1 and C_2 are made equal. Thus $R_1C_1 \doteq R_2C_2$ and the frequency of oscillation is given by

$$f_0 = \frac{1}{2\pi R_1C_1}$$

Now, the frequency is inversely proportional to the capacitance instead of the square root of the capacitance so that a 10-1 frequency range is obtained with variable capacitors of the type used in broadcast receivers. The frequency range of the oscillator is divided into bands. The change from one band to another is accomplished by the switching of R_1 and R_2 simultaneously so that R_1 is always equal to R_2. Continuous variation of frequency within the band is made possible by adjusting C_1 and C_2. These two capacitors are ganged so that their capacitances are always equal.

The cathode resistor of V_1 is a low-power tungsten lamp. Its purpose is to maintain the amplitude of oscillations constant over the operating range of frequencies. If the oscillations tend to increase, the current through the lamp increases. The filament becomes hotter and its resistance increases. Thus, there is a greater negative feedback voltage developed across the lamp. The increased degeneration reduces the gain of V_1 and holds the output from V_1 at nearly a constant amplitude.

15.6 Counting circuits. A *counting circuit* or *frequency divider* is a circuit which produces one output pulse for every two or more pulses applied to the input. Such circuits may be designed to operate reliably with frequency ratios from 1:2 to 1:10. Three types of counting circuits are in common use. The choice of circuit for a particular application is largely determined by the uniformity of frequency and amplitude of the input pulses. The circuits are discussed in the following sub-sections.

A. Scale-of-Two Counters

An Eccles-Jordan circuit requires two trigger pulses in the input circuit to produce one complete cycle of operation. By developing one output pulse from each cycle of operation, a 2:1 step down in frequency is achieved. Several Eccles-Jordan circuits connected in series may be used to produce output pulses at a submultiple of the input frequency. These circuits are referred to as scale-of-two, scale-of-four, scale-of-eight, etc. counters.

B. Ring Circuits

It is possible to connect several valves in series in a *ring circuit* with the output of each valve providing the input to the next in the ring. One valve is made to conduct current and all others are biased-off. When a trigger pulse is applied to all valves simultaneously, the conducting valve is turned off and the next valve in the series is turned on. The succeeding pulse turns off this latter valve and turns on the next in the ring. Thus if five valves are connected in a ring, any one valve responds to every fifth trigger pulse and a frequency division of 5:1 is achieved. One of these valves operates a counting relay which therefore is caused to respond to every fifth trigger pulse.

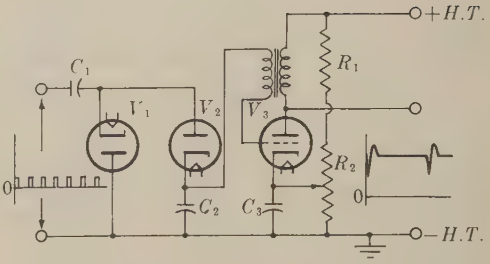


FIG.15-70 Step-by-step counting circuit.

C. Step-by-Step Counters

This type of circuit employs the principle that the p.d. between the plates of a capacitor increases in proportion to the amount of charge stored in the capacitor. Each input pulse is used to add one unit of charge to a storage capacitor. A definite number of these units is required to build up the p.d. on the capacitor to a predetermined value. When this value is reached, a following circuit is triggered which executes one cycle of operation and discharges the storage capacitor. A new series of charging steps then begins. Successful operation of this circuit requires that the amplitude of the input pulses remain at a constant value. The recurrence frequency of the pulses must not vary widely from the frequency for which the counter is adjusted.

A step-by-step counter is shown in Fig. 15-70. It consists of two diode valves V_1 and V_2 connected so that one conducts when the input voltage rises and the other when the input voltage falls. The input pulses which are usually square are applied to the diodes by way of a small differentiating capacitor C_1. A larger storage capacitor C_2 supplies a signal to the grid of V_3. This valve is connected in a blocking-oscillator circuit and is normally in a biased-off condition due to a positive potential applied to its cathode from the voltage divider R_1, R_2.

Positive pulses to be counted are applied to the input terminals. When the input voltage rises, C_1 charges up through V_2 and C_2. V_1 does not conduct during this time as its cathode is positive with respect to its anode. A charge is built up on C_2, however, and its voltage rises one step. When the input voltage falls at the end of the pulse, C_1 discharges through V_1. V_2 does not conduct during this period because its anode is made negative to its cathode. Since there is no path for discharge of C_2 during this time, it maintains its voltage. When the next input pulse is applied to the diode, an additional charge is added to C_2 by the charging of C_1. The p.d. across C_2 therefore increases another step. As this sequence continues, the potential on the upper plate of C_2 rises step by step. This potential is applied through the left winding of the pulse transformer to the grid of V_3. When this potential overcomes the bias potential imposed by the cathode, current begins to flow through the valve. A blocking-oscillator cycle takes place developing first a sharp drop and then a rise in anode potential providing the output as shown in Fig. 15-71. C_2 is discharged by grid current flow during the blocking-oscillator first half cycle and a new sequence of charging steps is begun. The counting ratio may be controlled by R_2 which sets the cathode bias for the valve. If the positive potential on the cathode is increased, more charging steps are required to cause the blocking oscillator cycle to take place. C_3 is a by-pass capacitor across R_2 and supplies the pulse of current required by the valve during its conducting period.

Fig. 15-71 shows the waveforms of the square voltage pulses applied to the input, the rising p.d. across the storage capacitor and the output voltage from the anode of V_3. The diagram shows the operation of the circuit adjusted to produce one output pulse from five input pulses. The voltage across the storage capacitor follows an exponential trend during the charging steps. Each succeeding step is smaller than the previous step because the input voltage must overcome the voltage across C_2 before V_2 conducts.

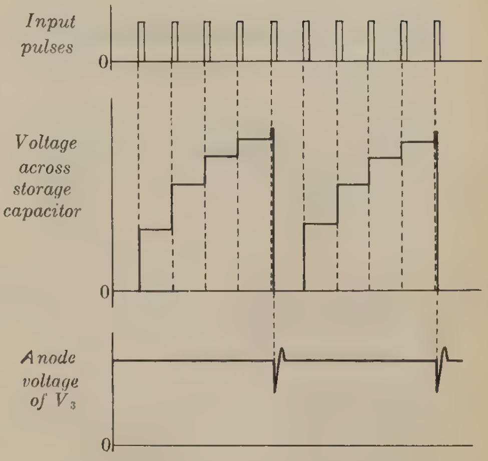


FIG. 15-71 Operation of step-by-step counting circuit.

Step-by-step counters operate most satisfactorily from pulses of constant recurrence frequency as well as constant amplitude. Leakage losses from the storage capacitor limit the time the capacitor can retain its charge between pulses. If part of the charge leaks off the storage capacitor between pulses, more charging steps are required to trigger the blocking oscillator. This may be compensated for by reducing the cathode bias provided the pulses arrive at a steady rate in the input.

15.7 Questions and problems.

1. Explain the purpose of a clipping circuit.
2. Draw a positive-clipping circuit using a series diode and explain its action with the aid of waveform diagrams.

3. Draw a diagram of a negative-clipping circuit using a parallel diode and explain its action illustrating the waveforms.

4. (a) Draw a diagram of a negative-clipping circuit using a biased diode.
   (b) Explain its action if the input voltage is a sine wave of 15 volts amplitude and the bias is obtained from a 10-volt battery.

5. Draw a circuit which will limit the amplitude of an input sine wave to 6 volts in both the positive and negative directions.

6. Draw a circuit which will clip an input sine wave of 15 volts amplitude and provide an output which consists of the negative peaks of the input varying between −12 volts and −15 volts.

7. (a) Show by means of a diagram how the circuit of question 6 can be modified to produce an a-c output which consists of negative pulses.
   (b) Explain the action of this circuit.

8. Draw a grid-clipping circuit and explain its action.

9. Draw a modification of the circuit of question 8 to produce increased grid clipping.

10. Explain the action of a saturation limiting circuit using diagrams.

11. Explain the action of a cut-off limiter using diagrams.

12. Draw the circuit of an over-driven amplifier and explain its action as a clipping circuit.

13. Explain the purpose of a clamping circuit drawing waveforms to illustrate its operation.

14. Explain, using diagrams of waveforms, how a clamping circuit differs from a clipping circuit.

15. (a) Draw a circuit of a diode clamper producing an output which is clamped to the negative side of ground potential.
    (b) Explain its action using waveform diagrams if the input is a square wave voltage varying between 25 volts and 75 volts.

16. Draw a circuit which will clamp a square-wave input voltage varying between 25 volts and 75 volts to the positive side of −15 volts.

17. Explain in detail the operation of the circuit of Fig. 15-22.

18. Explain in detail the operation of the circuit of Fig. 15-23.

19. Draw a grid-clamping circuit and explain its operation if the input voltage is a square wave varying between +10 volts and +30 volts.

20. Explain how different sawtooth sweep voltages of the same amplitude and frequency but with different sweep times can be applied to c.r.t. so that the trace starts from the same position on the screen for each sweep voltage.

21. (a) Draw the waveform of a trigger pulse voltage.
    (b) Explain the meaning of the terms pulse duration, pulse repetition rate and leading edge as applied to the trigger pulse voltage.

22. What is the purpose of a differentiating circuit?

23. (a) Draw the diagram of an *R-C* differentiating circuit.
    (b) Explain its action with diagrams if the input voltage is a square wave varying between 25 volts and 75 volts.

24. What is the purpose of an integrating circuit?
25. (a) Draw a diagram of an integrating circuit.
    (b) Describe its operation showing waveforms when a square wave voltage is applied which varies between +25 volts and +75 volts.
26. Draw the circuit of a shocked oscillator which produces a train of oscillations when a negative square pulse is applied and explain its operation.
27. Draw the circuit of a shocked oscillator which produces half cycles of oscillation when a square wave input voltage is applied and describe its operation.
28. Explain the purpose of a gate pulse.
29. What is a blocking oscillator?
30. What is a free-running blocking oscillator?
31. Explain the operation of a free-running blocking oscillator giving a diagram of the circuit and voltage waveforms.
32. What determines the duration of the quiescent period in the free-running blocking oscillator?
33. What is a synchronized blocking oscillator?
34. Draw the circuit for a synchronized blocking oscillator and explain its operation using waveforms to illustrate.
35. What determines the pulse duration in the synchronized blocking oscillator?
36. What determines the pulse repetition rate in the synchronized blocking oscillator?
37. Draw the circuit for a biased-off blocking oscillator and describe its action with the aid of diagrams.
38. Explain how the biased-off blocking oscillator may be triggered by a sine wave input voltage to produce one output cycle for every five cycles of sine wave input.
39. Draw the circuit of a thyratron pulse former and describe its operation.
40. Describe the charging action of the discharge line shown in Fig. 15-44.
41. Describe the discharging process of the discharge line of question 40.
42. Draw a circuit which uses a discharge line to produce large, sharp, output pulses when triggered by small input pulses.
43. Draw the circuit of a basic symmetrical multivibrator and explain its action by drawing the waveforms of the grid and anode voltages.
44. What determines the time of each multivibrator half cycle?
45. (a) What is the difference between a symmetrical and an unsymmetrical multivibrator.
    (b) Illustrate by drawing waveforms of the anode voltages.
46. What circuit changes are necessary in the circuit of question 43 to make it an unsymmetrical multivibrator?
47. Show by means of waveforms how the basic multivibrator may be synchronized to produce one cycle of output voltage for every eight input pulses.
48. Draw the circuit of a cathode-coupled multivibrator and explain its action using waveforms to illustrate.

49. Explain the difference in operation between the circuit of question 48 and the modified cathode-coupled multivibrator.

50. Draw the circuit of an electron-coupled multivibrator and explain its similarity to the basic multivibrator.

51. What is the advantage of the electron-coupled multivibrator over the basic multivibrator?

52. How is a biased-off multivibrator different from a free-running multivibrator in operation?

53. Draw the circuit of a biased-off multivibrator and explain its operation with waveforms.

54. Draw the circuit of a cathode-coupled biased-off multivibrator and the waveforms of the input trigger voltage and the output voltage.

55. Explain briefly the purpose of an Eccles-Jordan circuit.

56. Draw an Eccles-Jordan circuit and explain its operation using waveforms to illustrate.

57. Draw the circuit of a transitron square-wave generator and explain its operation giving waveform diagrams.

58. (a) Draw the circuit of a transitron sine-wave oscillator and explain its operation.

    (b) What circuit difference exists between it and the transitron square-wave oscillator?

59. What is the purpose of the Wien bridge in the Wien-bridge oscillator?

60. What are the conditions for oscillation?

61. Explain how the regenerative part of the feedback varies in amplitude.

62. Explain how the degenerative feedback is accomplished and why it remains constant.

63. How do the values of R_1 and R_2, C_1 and C_2 compare in a practical Wien-bridge oscillator circuit?

64. Give the formula for calculating the frequency of oscillation in the practical circuit of question 63.

65. Calculate the frequency of oscillation of the oscillator of Fig. 15-66 if $R_1 = R_2 = 500{,}000\Omega$, and $C_1 = C_2 = 300\mu\mu F$.

66. (a) What use may be made of a Wien-bridge oscillator?

    (b) In what respects is it superior to an *L-C* type of audio oscillator?

67. Explain the purpose of the tungsten-lamp resistor R_4 in the circuit of Fig. 15-66.

68. What is the purpose of a counting circuit?

69. Explain the action of a scale-of-four counting circuit.

70. Describe briefly the circuit arrangement and operation of a ring counting circuit producing a 4 to 1 step down in frequency.

71. Explain briefly the principle of the operation of the step-by-step counter.

72. Draw a step-by-step counting circuit and explain its action using waveforms to illustrate.

CHAPTER XVI

RADAR INDICATORS

The construction of the cathode-ray tube employing electrostatic deflection is described in Chapter I. The operation of this tube in the cathode-ray oscilloscope is discussed in Chapter IV. Applications of this instrument include visual observation of periodic voltage waveforms, and the determination of frequency and phase relationships between two voltages. Linear sweeps and sinusoidal sweeps are discussed in the description of the test oscilloscope. The present chapter deals more fully with cathode-ray tubes and the circuits associated with them for presenting information obtained by the receivers and transmitters of radar equipment.

16.1 Types of scan. There are many methods in use for displaying the information obtained by the radar set. The various types of scan or presentation are designated by letters, *A*, *B*, *C* up to *P*. Of these, the types *A*, *B* and *P* are most widely used in Service radar and are described in detail. The other types are, for the most part, modifications of these presentations used to display various combinations of range, bearing and elevation of targets.

A. The Type A Presentation

The *A* scan uses an electrostatic cathode-ray tube with a linear sweep applied to the horizontal-deflection plates and is used to measure the time which elapses between the transmission of a radar pulse and the return of its echo. This is accomplished by applying to the top *Y*-plate of the c.r.t. a positive voltage pulse at the instant the transmitter is turned on or *pulsed* and another at the instant the echo is received. The first of these pulses is synchronized with the start of the sweep and appears as a vertical upright *pip* at the left end of the trace. The echo pulse produces a second vertical pip farther along the trace. The distance between these two pips represents the total time taken for the transmitter pulse to travel to the target and the echo to return. Since the sweep is linear, a uniformly calibrated range scale may be fixed to the screen of the oscilloscope for measuring range directly. The zero mark of the scale is set opposite the transmitter pulse and the range is read from the scale opposite the echo pulse as indicated in Fig. 16-1. The duration of the sweep is determined by the desired range of the radar set. For example, if the 20-mile range scale is in use the time required for the sweep, which is the time required for a transmitter pulse to travel to a target 20 miles away and return, is 20 × 12.4 = 248 microseconds, since it takes 12.4 microseconds for a pulse to travel to a target one nautical mile away and return. There is a short waiting period before the next transmitter pulse is sent out and the sweep starts.

The bearing of the target is determined by the use of a highly directional aerial.

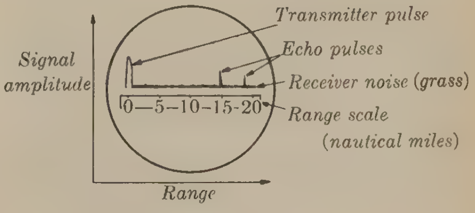


FIG.16-1 Type A presentation.

The amplitude of the echo pulse is observed as the aerial is rotated. When the echo pulse reaches a maximum amplitude on the c.r.t., the aerial is pointing in the direction of the target, and its bearing may be noted. This type of scan is widely used in surface-warning radar sets.

B. The Type B Presentation

A type *B* scan presents both range and bearing information of the target on the c.r.t. This type of presentation is illustrated in Fig. 16-2. The targets appear as bright spots on the screen. The position of the spot above the base line is a measure of the distance between the radar set and the target. The displacement of the spot from the vertical centre line indicates the bearing of the target relative to the ship's head. Target 1 is dead ahead at 15 miles, target 2 is 70° port at 10 miles and target 3 is 45° starboard at 4 miles distance.

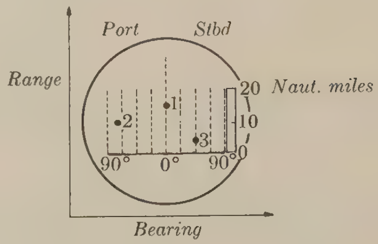


Fig.16-2 Type B presentation.

This presentation is effected by producing a linear sweep in a vertical direction on the screen beginning at the base line when the transmitter is pulsed. The entire sweep is shifted in a horizontal direction, synchronized with the bearing of the transmitting and receiving aerial. For example the first sweep is started at the 90° position at the left end of the base line when the aerial is pointing 90° port. The transmitter is pulsed and the sweep started simultaneously at this point. As the transmitter pulse travels to the target and the echo returns, the sweep progresses vertically upward from the starting point. If an echo is received during this sweep, it is applied as a positive pulse to the grid of the c.r.t. momentarily. The electron beam, which is normally biased off, is allowed to pass through the c.r.t. and strikes the screen at a point which therefore indicates the range and bearing of the target. When this sweep is completed, the electron beam is re-directed to the base line. It is then shifted 1° to the right, or to the position designated as 89° port. At the same time the aerial is re-directed from 90° to 89° port. A new sequence of transmitter pulsing and indicator sweeping then takes place. The entire rectangular area on the screen is scanned very rapidly as the aerial sweeps from 90° port to 90° starboard. The aerial is switched back to 90° port and a new series of sweeps initiated on the left side of the c.r.t. Screens having long persistence of vision are used in this type of presentation. The target pulses are presented with sufficient rapidity that the pattern fades only slightly between scans.

C. The Type P or PPI Presentation

A *plan-position indicator* (*PPI or type P*) scan is a bearing and range presentation of all targets which lie around the radar set within its range. The radar set position occupies the centre of the screen. Targets due north of the set appear as bright spots along a line from the centre to the top of the c.r.t. The range to the target is indicated by the distance from the centre of the screen to the bright spot. A typical *PPI* presentation is shown in Fig. 16-3. A linear sweep is started from the centre of the screen each time the transmitter is pulsed. The sweep proceeds radially from the centre as the pulse travels out from the radar set and the echo returns. The direction of the sweep corresponds to

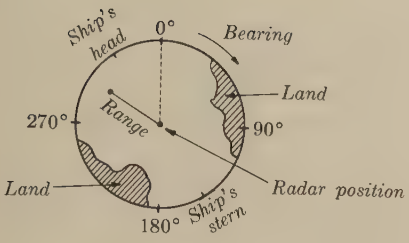


FIG.16-3 Plan-position-indicator (PPI) presentation.

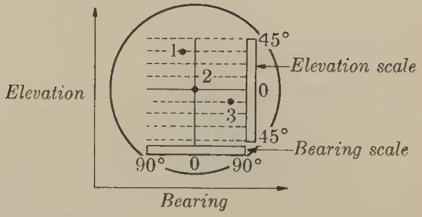


FIG.16-4 Type C presentation.

the direction in which the aerial is pointing during that sweep. The vertical sweep above the centre of the tube takes place when the aerial is pointing due north. If an echo is received, it is applied as a positive pulse of voltage to the grid of the tube. It overcomes the normally biased-off condition of the electron beam and causes a bright spot to appear on the screen. Land echoes appear as large patches of brightness on the front of the tube. The aerial is made to rotate continuously through 360° as the sweep rotates in the *PPI* tube. The entire presentation is repeated with sufficient rapidity that the pattern barely fades out between successive sweeps. This type of presentation is of particular value as a general viewing scan. It has the advantage over the other types that all targets within range and with any bearing may be kept under observation at one time.

D. Other Types of Presentation

While the type *A*, type *B* and *PPI* scans are the most widely used presentations in Service radar equipments, other types of scans are sometimes used to meet particular requirements; these types are lettered *C*, *D*, etc. Reference is made here to the type *C* presentation which is a plot of elevation against bearing. The circuits used to produce the sweeps are described in later sections of this chapter.

A *C* scan presentation is shown in Fig. 16-4. Since range is not given by this presentation, the linear time base employed in the *A* and *B* scans is not used as such. At any instant the electron beam is directed to a particular spot on the c.r.t. screen which corresponds to the direction in which the aerial is pointing relative to the line of flight of the aircraft containing the radar set. The transmitter is pulsed and a period of time allowed to elapse for the return of the echo. During this time the electron beam is directed to a fixed spot on the screen, but the tube is biased off. If an echo is received, the bias is removed momentarily and a bright spot is produced on the screen. The active area of the c.r.t. is scanned as the aerial scans the area bounded by 90° port to 90° starboard and 45° above to 45° below the line of flight. When the antenna is directed 90° port and 45° upward, the beam is directed to the upper left corner of the c.r.t. For the next transmitter pulse, the antenna is shifted from 90° to 89° port and the beam is re-directed one unit to the right. When the top line of the c.r.t. has been covered, the angle of elevation of the antenna is reduced and the electron beam is dropped a corresponding amount in the c.r.t. Target 1 is 20° to port and 40° above the line of flight. Target 2 is dead ahead while target 3 is 80° to starboard and 15° below the line of flight.

16.2 Linear sweeps for electrostatic-deflection tubes. A linear sweep is one

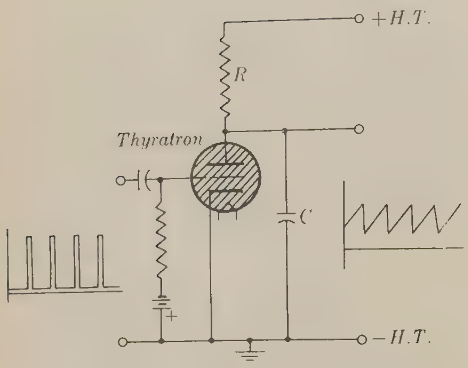


Fig.16-5 Thyratron sweep generator circuit.

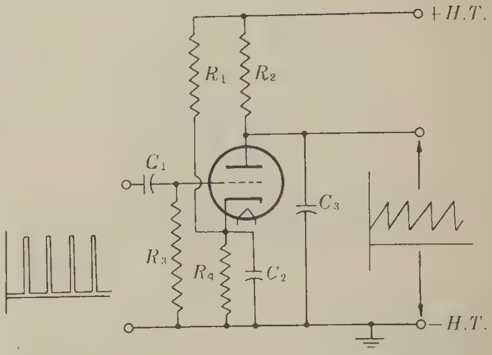


Fig.16-6 Sweep generator using high-vacuum triode.

in which the electron beam is deflected from one side of the c.r.t. to the other at a uniform rate. This causes the spot of light to move equal distances across the screen in equal times. A fixed scale attached to the front of the tube is calibrated to read the distance between radar set and target. It is important that the sweep be linear in order that equal spaces on the scale may correspond to equal distances traversed by the radar pulse. This requires that the deflection voltages applied to the X-plates rise or fall at a uniform rate during the sweep.

A. Oscilloscope Sweep Generators

A linear sweep voltage generator used in a test oscilloscope is described in Chapter IV. The sweep voltage is obtained from a capacitor C by permitting it to charge through a resistor R (Fig. 16-5). A thyratron valve is connected in parallel with the sweep capacitor to act as a switch. While the valve is biased off due to the bias battery it acts as an open circuit and permits the capacitor to charge. When the valve is caused to conduct by a sync. pulse it acts as a short circuit discharging the capacitor quickly. The voltage across the valve drops to the de-ionization potential, the valve stops conducting and C begins to charge again. The RC time constant is being compared to the time between sync. pulses. A sawtooth voltage is developed across the capacitor as indicated by the waveform in Fig. 16-5. The sweep voltage is actually part of an exponential rise toward the potential of the charging source. The sweep is maintained fairly linear by employing only 10% or less of this complete voltage rise. Alternative sweep generating circuits are described in the following paragraphs. Various methods of improving linearity are suggested.

In the sweep generator circuit of Fig. 16-6 a vacuum triode is used instead of a thyratron. The sweep capacitor C_3 is connected in parallel with the valve and its cathode circuit. The valve is normally biased beyond cut-off by a positive potential on its cathode from R_1, R_4. Since no current flows through the valve in this condition it acts as an open switch. C_3 therefore begins to charge exponentially through R_2 toward the supply voltage. When a small fraction of this voltage has been built up on C_3, however, a positive voltage pulse is applied to the grid of sufficient amplitude to overcome the bias and make the valve conduct heavily. C_3 discharges through R_4C_2 and the valve. The anode potential and sweep voltage fall to a low value in a short time producing the fly-back portion of the sweep voltage. The positive pulse is then removed from the grid

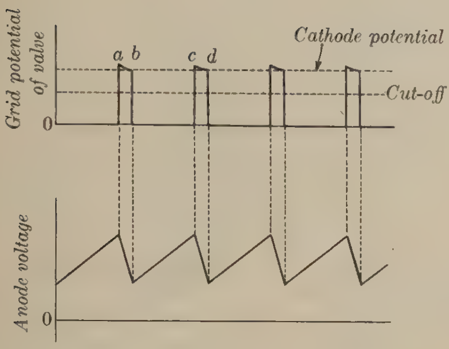


FIG.16-7 Voltage waveforms of a triode sweep generator.

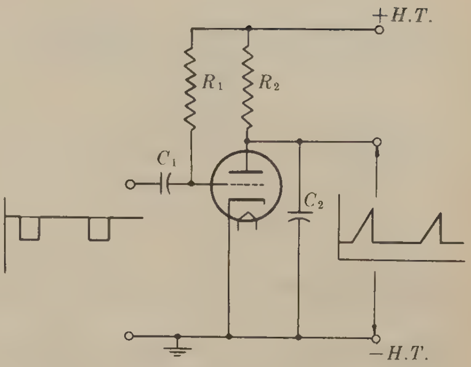


FIG.16-8 Triode cut-off sweep generator.

and the valve again stops conducting. C_3 again begins to charge through R_2 to produce another forward sweep. The p.d. across C_3 is allowed to rise and fall only a small fraction of the supply voltage so that the sweep is practically linear. C_2 is a by-pass capacitor across R_4 to maintain a steady bias voltage across R_4. Waveforms of the input and output voltages are shown in Fig. 16-7. The periods *ab* and *cd* represent the times when the valve is made to conduct and the fly-back voltage is developed. The period *bc* represents the time when the valve is cut off and the sweep voltage is produced.

If the duration of the positive voltage to the grid is extended, the valve conducts for a longer time than is required to discharge C_3. The anode potential remains low during this time and the next sweep is delayed until the valve is again cut off. It is frequently necessary in radar circuits to allow a waiting period between the fly-back of one sweep and the beginning of the next sweep. The circuit just described is not the best type for this purpose as considerable driving power is required to maintain the grid positive in spite of grid current flow. The following circuit is better suited to this type of operation.

B. Triode Cut-off Sweep Generator

The circuit of Fig. 16-8 differs from the preceding circuit in that the valve is normally made to conduct heavily instead of being cut off. This is accomplished by connecting the grid to +H.T. through a large resistor R_1. The flow of grid current through R_1 develops a large p.d. across R_1 which is almost equal to the full supply voltage. The grid therefore is only slightly positive to the cathode, the amount being just sufficient to maintain the flow of grid current.

A large anode current also flows developing a large p.d. across R_2. The anode potential is low and therefore the p.d. across the sweep capacitor is small. The input voltage to the grid of the valve is a negative square wave beginning when the sweep is to commence. This input must be of sufficient amplitude to drive the grid beyond cut-off, and must last for the duration of the sweep. The valve therefore becomes an open circuit and C_2 begins to charge through R_2 exponentially. When the negative voltage on the grid ends, the valve conducts heavily again. The sweep capacitor discharges quickly through the valve to its original p.d. This sudden fall in voltage across C_2 produces the back sweep in the c.r.t. A waiting

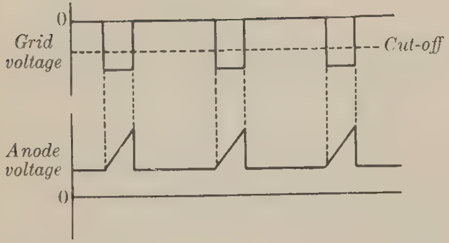


FIG.16-9 Voltage waveforms of triode cut-off sweep generator.

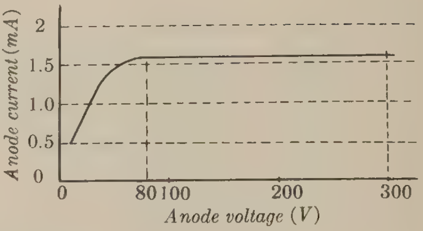


FIG.16-10 $e_b - i_b$ characteristic of a pentode valve.

period then follows until the grid is again driven negative by the input negative square wave voltage. The waveforms of input and output voltages are shown in Fig. 16-9. If a shorter sweep time is desired, a negative square wave of shorter duration is applied to the grid. At the same time, in order to maintain the same sweep voltage amplitude, and hence the same length of trace, the C_2R_2 time constant must be made smaller. This means that C_2 can be made to charge to the same p.d. in the shorter time.

C. Constant-Current-Pentode Sweep Generator

In the preceding sweep circuits the sweep voltage is provided by the charge of a capacitor through a resistor. This provides an exponential rise in voltage. The sweep voltage, however, is made fairly linear by using only a small part of the exponential rise. The sweep capacitor may be made to charge or discharge through a valve instead of a resistor. If the charging current of a capacitor is maintained constant, the amount of charge on the capacitor must increase at a uniform rate. Consequently the p.d. across the capacitor must increase at a steady rate. Similarly, if the discharge current of a capacitor is kept constant, its p.d. must fall linearly. The discharge current may be maintained constant by allowing the capacitor to discharge through a pentode valve of the constant-current type. The $e_b - i_b$ characteristic curve of a sharp cut-off pentode valve is shown in Fig. 16-10. The anode current is observed to be practically constant for any anode voltage in the range from 80 volts to 300 volts. The anode-to-cathode d-c resistance of the valve is proportional to the anode voltage in this region of operation. A constant-current-pentode sweep generator is shown in Fig. 16-11. The pentode V_2 is connected in series with a triode valve V_1 across the supply voltage. V_1 is made to conduct heavily with only a small part of the H.T. voltage between anode and cathode. This is accomplished by connecting the grid of the valve to +H.T. through a resistor R_2. The flow of grid current through R_2 develops a p.d. across it which limits the grid potential to a value which is slightly more positive than the cathode.

The grid of the pentode valve on the other hand is biased somewhat negatively with respect to its cathode. The cathode current flowing through R_3 maintains the cathode at a positive potential while the grid is connected to ground. The sweep capacitor C_3 is connected in parallel with V_2. The screen dropping resistor R_1 and by-pass capacitor C_2 complete the circuit. Most of the supply voltage appears between cathode and anode of V_2 and the anode potential of this valve is high. The sweep capacitor is charged to the anode potential of V_2.

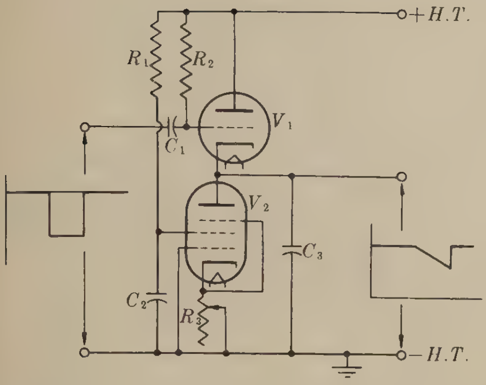


FIG.16-11 Constant-current-pentode sweep generator.

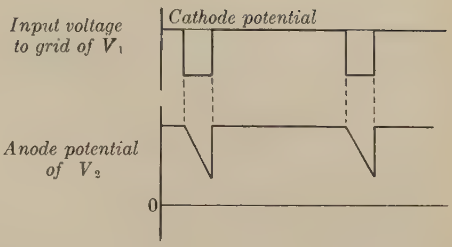


FIG.16-12 Voltage waveforms of constant-current-pentode sweep generator.

The input to the circuit is a negative square wave of voltage of large amplitude applied via C_1 across R_2 to the grid of V_1 at the instant when the sweep is to begin in the c.r.t. (Fig. 16-12). This large negative potential at the grid of V_1 cuts off the current through this valve. This in effect, disconnects V_2 and C_3 from the H.T. supply. C_3 begins to discharge through R_3 and V_2. As C_3 discharges, the current through V_2 remains constant and the p.d. across C_3 decreases linearly as indicated in Fig. 16-12. The fall in anode potential is limited to the straight portion of the valve's characteristic curve in order to maintain a constant discharge current. At the end of the sweep, the negative voltage on the grid of V_1 returns suddenly to normal. This valve conducts again and its cathode potential rises. C_3 charges rapidly through V_1 producing the backsweep voltage. The output from V_2 is therefore a negative-going sweep voltage which lasts for the duration of the negative voltage to the grid of V_1. The rate of discharge of C_3 is determined by the value of capacitance used and the resistance of R_3 and the pentode. The variable resistor R_3 controls the bias on the valve and hence the magnitude of the discharge current. Thus R_3 varies the amplitude of the sweep voltage. The cathode by-pass capacitor is omitted from the circuit to introduce a measure of degeneration and improve linearity. If the discharge current tends to decrease during the sweep, the p.d. across R_3 tends to decrease. The bias on the valve therefore becomes less, which tends to increase the discharge current to its original value.

D. Cathode Follower Feedback Circuit

The linearity of the triode cut-off sweep discussed in *B*, may be considerably improved by the use of a cathode follower circuit with feedback to the sweep generator. Such a circuit is shown in Fig. 16-13. V_1 is the sweep generator. This valve normally conducts heavily because of its positive grid condition. Its anode potential and therefore the voltage across the sweep capacitor are low. The grid of V_2 is directly coupled to the anode of V_1. V_2 in the cathode follower circuit conducts a small amount of current which raises its cathode potential above its positive grid potential. The feedback line is from the cathode of V_2 through C_2 to the junction of R_2 and R_3 in the anode circuit of V_1.

When the input negative square wave voltage to the grid of V_1 cuts off the current through this valve, C_3 begins to charge through R_3 and R_2 in the normal manner. The positive-going sawtooth voltage across C_3 is directly coupled to the grid of V_2. As the potential of this

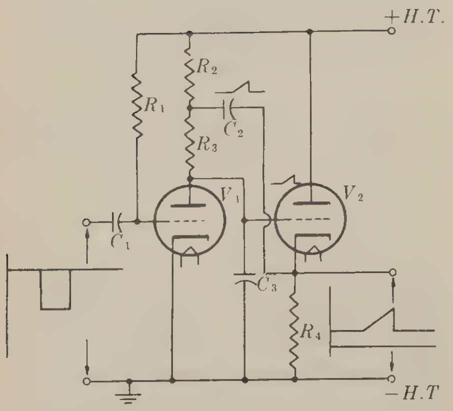


FIG. 16-13 Cathode follower feedback circuit to improve linearity.

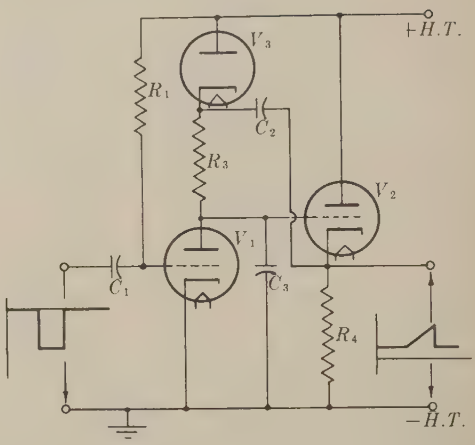


FIG. 16-14 Cathode follower feedback circuit using a diode.

point rises, the potential of the cathode also rises. This rising voltage is fed back through C_2 to the upper end of R_3. The potential of this point rises as the sweep voltage increases. This effectively increases the supply voltage to the charging capacitor as the sweep progresses. Since the output voltage of a cathode follower is almost as large in amplitude as its grid voltage, the upper end of R_3 rises in potential almost as rapidly as the lower end. Therefore a constant current flows through R_3 and the sweep capacitor charges at a uniform rate. There is no voltage gain in the cathode follower circuit. Its purpose is solely to improve linearity.

The operation of the circuit departs slightly from the ideal for three reasons. The output voltage of the cathode follower is never quite as large as its input. Hence the feedback voltage to the top of R_3 is always slightly less than the rise in voltage at the anode of V_1. Also C_2 must charge slightly during the sweep so that the rise in potential on its left plate is a little less than the rise on the right plate. This loss in voltage is kept small by making C_2 large. Finally, R_2 acts as a load on the feedback line in parallel with R_3 and the charging circuit. Its effect may be made small by making R_2 large in comparison with R_3. A better arrangement is to replace R_2 by a diode valve with its anode connected to +H.T. and its cathode to R_3 and C_2 (Fig. 16-14).

When V_1 is conducting V_3 also conducts and since its impedance is very low the potential of the upper end of R_3 is nearly +H.T. When the negative square wave is applied to the grid of V_1 this valve is cut off. C_3 begins to charge through R_3. The rising voltage across C_3 is applied to the grid of V_2 and causes a rise in voltage across R_4. This rise in voltage is coupled via C_2, which has a large capacitance, to the upper end of R_3. The upper end of R_3 rises above +H.T. potential and V_3 is cut off. Since the diode acts as an open circuit the potential at the top of R_3 may rise well above the H.T. voltage. This action improves the linearity of the sweep voltage and also a larger sweep amplitude is obtained with a given power supply voltage.

E. Sweep Integrating Circuit

Another method of improving the linearity of a sweep voltage is shown in the circuit of Fig. 16-15. The sweep

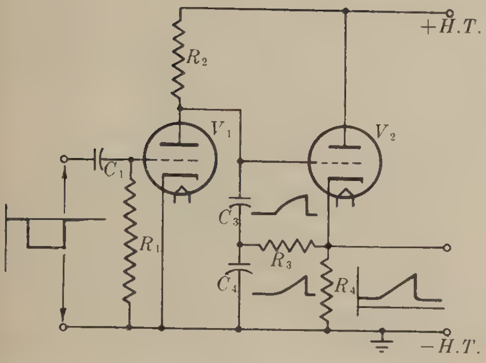


FIG. 16-15 Sweep integrating circuit to improve linearity.

voltage is developed across two capacitors C_3 and C_4 in series. R_3 and C_4 comprise an integrating circuit from the cathode of V_2 to the lower side of C_3. The waveforms of voltages across C_3, C_4 and R_4 are shown in Fig. 16-16. The voltage across C_3 rises rapidly at first but levels off somewhat as the charging current through R_2 decreases exponentially. This is the source of non-linearity. The rising voltage at the grid provides a rising voltage at the cathode of V_2 which is applied to the integrating circuit R_3C_4. This develops a voltage across C_4 which rises slowly at first and more rapidly as its applied voltage increases. The combined effect of these two rising voltages is applied to the grid of V_2. If the value of R_3 is correctly chosen, the non-linearity of the sweep generator may be offset by the integrating effect of R_3C_4 and a linear sweep voltage results.

Thus advantages of feedback from the cathode of V_2 to the anode circuit of V_1 may be combined with that of integration just described to provide a sweep voltage of good linearity and large amplitude without the use of an excessively high value of supply voltage.

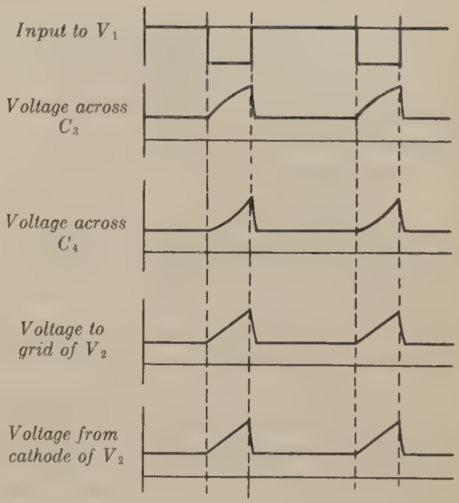


FIG. 16-16 Voltage waveforms of sweep integrating circuit.

F. Miller Time-Base Generator

The Miller time-base generator (Fig. 16-17) employs a pentode valve with the sweep capacitor C_2 connected between anode and grid. In its steady state condition electrons flow in the valve from the cathode to the control grid and also to the screen grid. The grid current flowing through R_2 develops a large p.d. which maintains the grid potential at one volt, or less, positive to the cathode. The screen potential is also low, due to the flow of screen current. No anode current flows during this time as the suppressor is maintained below the cut-off potential by the input square wave voltage. The anode potential is therefore at H.T. and the sweep capacitor C_2 is charged to almost the full H.T. voltage.

When the sweep starts, the suppressor potential is raised to zero by the square wave of input voltage. This sudden rise in suppressor potential causes a sudden drop in anode potential from *a* to *b* (Fig. 16-18). In the absence of C_2, a large anode current would begin to flow and the anode potential would fall suddenly to a very low value. A negative square wave of voltage would then be applied across C_2R_2 to the grid of the valve as shown in Fig. 16-18 (b). This would result in an exponential rise in

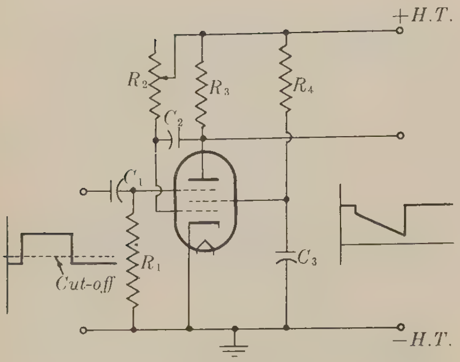


FIG. 16-17 Miller time-base generator circuit.

grid voltage, Fig. 16-18 (c), and an exponential fall in anode voltage, Fig. 16-18 (d). But the rising grid potential point and the falling anode potential point are connected by C_2. Any non-linearity in the grid voltage is amplified by the valve and fed back through C_2 in anti-phase. That is, a sudden rise in grid voltage would produce a sudden fall in anode voltage of greater magnitude. This voltage, fed back through C_2 would counteract the initial rise in grid voltage assumed at the start. As a result the grid potential rises linearly as shown in Fig. 16-18 (e) and the anode potential falls linearly, Fig. 16-18 (f).

This linear portion of the anode voltage waveform is used as the sweep voltage. The sweep is ended by the input voltage when it drops the suppressor potential below cut-off again. C_2 recharges through the grid-cathode circuit of the valve and R_3. A valve with high gain is used in the Miller time-base generator to maintain a high degree of linearity.

The rate at which the anode voltage falls is determined by the values of C_2 and R_2 as well as the voltage applied to the upper end of R_2. An increase in the values of C_2 or R_2 causes a decrease in the rate at which the sweep voltage falls. An increase in potential at the upper end of R_2 causes an increase in the speed of the sweep.

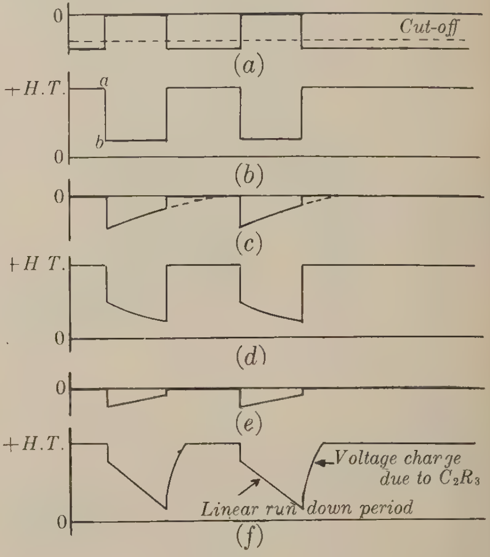


FIG. 16-18 Voltage waveforms of Miller time-base generator.

In addition to its advantage of producing a sweep voltage of improved linearity, the Miller time-base circuit is a useful circuit for producing a sweep of long duration. This results from the Miller effect described in Chapter II in which it is shown that the input capacitance of the valve is given by the formula

$$C_i = C_{gk}+C_{ga}\,(1+A).$$

This formula indicates that there is an effective amplification of capacitance in the grid-cathode circuit. For example, a circuit with a gain of 50, having a small external capacitance connected between grid and anode, operates as though it had a capacitance 51 times as great connected between grid and cathode. In the case of a multi-stage amplifier containing an odd number of stages, a capacitor of small capacitance connected between the grid of the first stage and the anode of the last stage may be used in place of a capacitor several thousand times as large between grid and cathode. In this way a sweep duration of several minutes is obtained without the use of a prohibitively large sweep capacitor in the grid circuit.

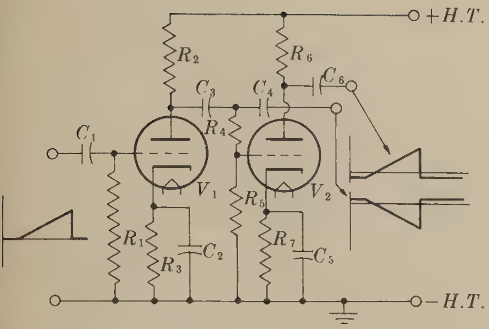


FIG. 16-19 Sweep-voltage phase inverter.

16.3 Sweep-voltage amplifiers. The advantages of balanced sweep circuits for cathode-ray tube deflection are discussed in Chapter IV. The inherent de-focusing effect of the unbalanced circuit is made considerably less by using a balanced circuit. Also, the amplitude of voltage applied to each deflecting plate must be only half of the amplitude necessary in an unbalanced circuit. Balanced deflection requires that two voltages of equal amplitudes and 180 degrees out of phase with each other be applied to the X-plates. While the right X-plate is being made more positive to attract the electron beam, the left X-plate must be made progressively more negative to repel the beam. Several circuits used to deliver these balanced sawtooth voltages are described in the following paragraphs. They are called *inverters* or *paraphase amplifiers*.

A. Phase-Inverter Circuit

Reference to a resistance-capacitance coupled amplifier for inverting the phase of a voltage waveform is made in Chapter II. Such a circuit is shown in Fig. 16-19. It consists of two class A voltage amplifiers so that the wide range of frequencies contained in the sawtooth waveform will be amplified uniformly. R_1 and R_5 are the grid-leak resistors and R_2 and R_6 the anode load resistors. C_1, C_3, C_4 and C_6 are coupling capacitors. Cathode bias is provided by C_2R_3 for V_1 and by C_5R_7 for V_2.

The input voltage to V_1 is shown here as a positive-going sweep voltage applied to the grid. Hence the output from the anode of V_1 is a negative-going sweep voltage of increased amplitude. It is coupled through C_3 and C_4 to the left X-plate of the c.r.t. This negative-going voltage is also coupled through the voltage divider R_4, R_5 to the grid of V_2. The values of R_4 and R_5 are so chosen that voltage changes at the anode of V_1 are reduced at the grid of V_2 by a factor equal to the amplification of V_2. Thus if V_2 has a gain of 10, the grid voltage changes to this valve are made one-tenth of the anode voltage changes of V_1. The output voltage from V_2 therefore has the same amplitude as that from V_1. The positive-going sweep voltage from the anode of V_2 is coupled through C_6 to the right X-plate of the c.r.t.

In order to prevent distortion of the waveform at the grid of V_2 it is necessary to compensate for the grid-cathode capacitance of this valve. For the low-frequency components of the sawtooth waveform, the reactance of the grid-cathode capacitance is negligible. The voltage division is adequately provided by the resistors R_4 and R_5. For the high-frequency components, this capacitance provides a low impedance across R_5 which interferes with the voltage division. This may be compensated for by placing a capacitor C_8 in parallel with R_4 as shown in Fig. 16-20. The capacitance is

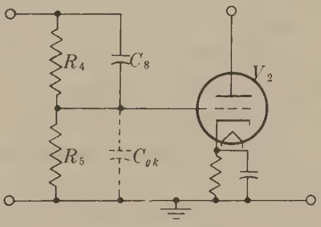


FIG. 16-20 High-frequency compensation in sweep phase inverter.

of such a value that the reactance of C_8 and of the grid-cathode capacitance are in the same ratio as the resistances of R_4 and R_5. Hence the same voltage division is maintained at all frequencies and this source of distortion is eliminated.

B. Split-Load Amplifier

The split-load amplifier of Fig. 16-21 is a one-valve paraphase amplifier that provides equal out-of-phase voltages of amplitudes slightly less than the input. It contains two equal load resistors R_2 in the anode circuit and R_3 in the cathode circuit. When the input voltage rises, the current through the valve increases. This produces an increasing p.d. across both R_2 and R_3. The anode voltage falls as the cathode voltage rises. Since the resistances of R_2 and R_3 are equal, the changes in voltage across them must also be equal. The cathode circuit of this amplifier operates as a cathode follower. Consequently the amplitude of the output voltages is always slightly less than the amplitude of the input voltage.

C. Self-Balancing Phase Inverter

While the circuits of Fig. 16-19 and Fig. 16-21 are simple in their basic operation, they may introduce distortions which prohibit their use in accurate ranging indicators. Distortion is introduced chiefly by the loss of high-frequency components from the voltage waveform. Any distortion introduced by the first valve is amplified by the second valve circuit. More elaborate circuits have been designed to keep these unwanted effects to a minimum. One such circuit is shown in Fig. 16-22. The phase inverter V_3 is ordinarily used in conjunction with the sweep generator and the cathode follower feedback circuit of Fig. 16-13 as shown in Fig. 16-22. The positive-going voltage from the cathode of V_2 is coupled through C_6 to the right X-plate of the c.r.t. Also, this voltage is applied by way of a voltage-dividing network consisting of C_4, C_5, R_5 and R_6 to the grid of V_3. This valve operates as a Class A amplifier which inverts and amplifies the signal. A negative-going sweep voltage is coupled through C_7 to the left X-plate of the c.r.t.

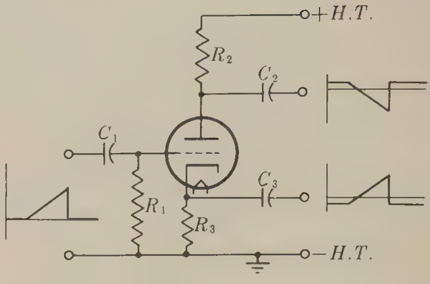


Fig. 16-21 Split-load sweep amplifier.

The operation of the voltage-dividing network is most readily understood by considering a particular example. The positive-going sweep voltage at the cathode of V_2 has an amplitude of 150 volts. The negative-going sweep voltage at the anode of V_3 must also have an amplitude of 150 volts. If the gain of the V_3 amplifier is 10, the amplitude of the voltage applied to the grid of V_3 must be 15 volts. This voltage is obtained for low frequencies from the voltage divider R_5, R_6. The high-frequency voltage division is provided by the capacitors C_4 and C_5. During the sweep, the potential at the left end of C_4 rises 150 volts, the potential at the right end of C_5 falls 150 volts and the potential between C_4 and C_5, that is, the V_3 grid potential falls 15 volts. The voltage change across C_4 is therefore $150-15$ or 135 volts. The voltage change across C_5 is $150+15$ or 165 volts. The capacitances of C_1 and C_2 must therefore be in the inverse ratio of 135 to 165. Thus the capacitances might be $C_1 = 330\mu\mu f$ and $C_2 = 270\mu\mu f$.

The reason for connecting C_5 between anode and grid of V_3 is to provide a

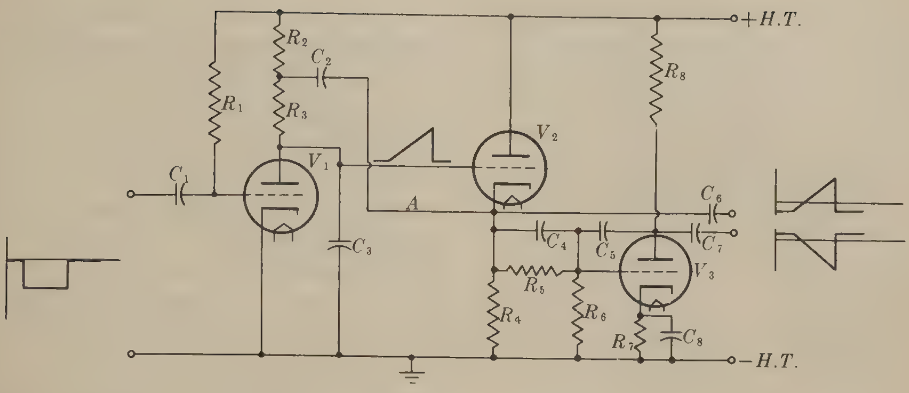


FIG. 16-22 Self-balancing phase inverter following a cathode-follower-feedback sweep generator.

self-balancing action that tends to maintain the output voltages of V_2 and V_3 of equal amplitude despite changes in the amplification of V_3. Thus, not only do C_4 and C_5 provide a high-frequency voltage dividing network but also they provide the self-balancing action for all frequencies. The self-balancing action is such that if the output of V_3 decreases due to change in amplification of V_3 the input of V_3 is caused to increase, thus opposing the decrease of the output. As a result a very large percentage change in the amplification of V_3 must take place before the output of V_3 changes by a small percentage. For example, suppose the output voltage of V_3 decreases by 5% due to a change in amplification. Then the amplitude of this voltage falls by 5% of 150 or 7.5 volts. The new amplitude of output is then 142.5 volts. The complete change in potential across C_4 and C_5 is now $150+142.5$ or 292.5 volts. The change in voltage across C_4 is $\frac{270}{270+330} \times 292.5$ or 132 volts. Therefore the rise in voltage on the grid of $V_3 = 150-132$ or 18 volts, as compared with 15 volts for normal operation. The amplification of this valve is then $\frac{142.5}{18}$ or 7.92 instead of the original value of 10. The amplification therefore must have decreased by $\frac{2.08}{10}$ or 20.8 percent. Thus, a change in amplification of 20.8% causes a change in output voltage of only 5%. The balancing network therefore serves to equalize the amplitudes of the sweep voltages. Moreover this equalizing effect is produced at every point along the sawtooth waveform. In this way distortion of the linear waveform by non-linear amplification of V_2 is greatly reduced.

D. Cathode-Coupled Phase Inverter

The paraphase amplifier shown in Fig. 16-23 consists of two degenerative amplifiers coupled together by a common cathode resistor R_6. Each valve operates with its grid slightly negative to its cathode. Bias cannot be obtained from the cathode circuit because a varying cathode potential is required to provide a signal to V_2. A negative d-c potential is applied to the grid of V_1 from a bias battery or equivalent source. The grid of V_2 is operated at a steady positive potential from the voltage divider R_3, R_4 but is less positive than the cathode. R_2 and R_5 are the anode load resistors of the two valves.

A positive-going sweep voltage is applied to the grid of V_1 via the capacitor

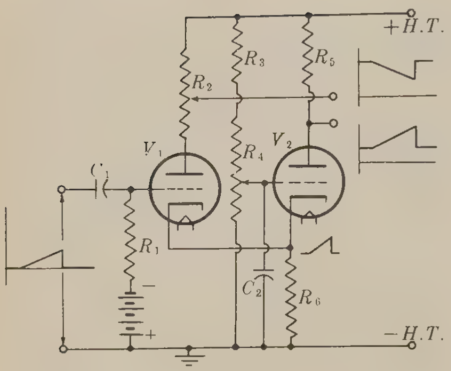


FIG. 16-23 Cathode-coupled phase inverter.

C_1. A negative-going sweep voltage is therefore developed at the anode of V_1 which is applied to the left X-plate of the c.r.t. The increasing current through R_6 causes the cathode potential to rise. This is equivalent to a fall in grid potential for V_2. Therefore, the output voltage from V_2 is the positive-going sweep voltage which is applied to the right X-plate. C_2 is a by-pass capacitor which maintains a steady voltage on the grid of V_2 throughout the cycle.

The degeneration introduced by the varying cathode potential reduces the gain of each stage to about one-half of the gain that V_1 would have with fixed bias. This is illustrated by the following example. Suppose a rise in potential of 20 volts is applied to the grid of V_1 and that the value of R_6 is such that the increase in current causes the cathode potential to rise 10 volts. Thus the grid of V_1 has been raised 10 volts with respect to its cathode while the cathode potential of V_2 has been raised 10 volts with respect to its grid. The effective grid-cathode signal of each valve is therefore one-half of the input signal to V_1. Thus the output voltages of V_1 and V_2 each have an amplitude of one-half the output of a single tube operating with a fixed bias. The picture is further complicated by the fact that the amplifications and anode load resistors of the two valves must be unbalanced. If this were not the case, the current through V_1 would increase by the same amount as the current through V_2 decreases. There would be no change in current through the cathode resistor and no change in cathode potential. Consequently there would be no signal voltage applied to V_2. The anode current change in V_1 must be greater than the anode current change in V_2. In order to obtain equal outputs from the two anodes, the anode load resistor of V_1 must be smaller than that of V_2. Alternatively, equal load resistors may be employed and only a part of the larger output of voltage used, as indicated by the variable tap on R_2.

This type of phase inverter is sometimes directly connected to the horizontal-deflection plates of the c.r.t. The circuit must then supply d-c centering voltages as well as the sawtooth sweep voltages. Centering is provided by R_4 which permits adjustment of the d-c grid potential of V_2. For example, if this grid is made more positive by adjusting R_4 an increased current flows through V_2 during the period between sweeps. The cathode potential is therefore higher and less current flows through V_1. Under these conditions the anode potential of V_1 would be higher and the anode potential of V_2 would be lower during the period between sweeps. The trace would then be centered farther to the left on the screen of the c.r.t.

16.4 Electromagnetic deflection circuits. A description of the structure and basic operation of a cathode-ray tube with electromagnetic deflection is given in Chapter I. If the electron beam from the gun of the tube is made to pass through a magnetic field perpendicular to the axis of the tube, the beam is deflected. The direction of deflection is at right angles to both the direction of the electron path and the

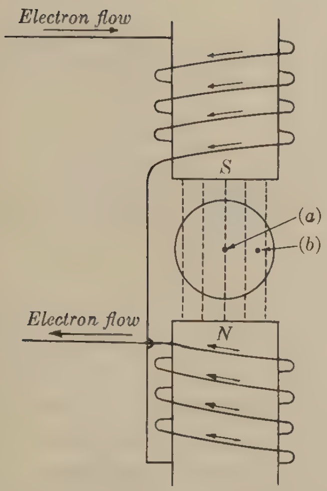


FIG.16-24 Electromagnetic deflection of the electron beam.

direction of the magnetic field. Since an electron beam constitutes an electric current, the direction of deflection is given by the motor principle. This is illustrated in Fig. 16-24. The circle represents the neck of the c.r.t. in cross-section. The electron beam from the electron gun is approaching the observer in a direction perpendicular to the plane of the paper. The neck of the tube passes between the poles of the electromagnet. When no current flows through the deflection coils, there is no magnetic field and the electron beam passes through the centre of the tube as indicated by (a). When electrons flow through the coils in the direction shown, the upper pole of the soft-iron core becomes S and the lower pole N. The magnetic lines of force are directed upward in the region between the poles and the electron beam is deflected to the right as indicated by (b). The amount of deflection is proportional to the field strength and hence to the amplitude of the current flowing in the coils. If the current is reversed, the magnetic field is reversed and the beam is deflected toward the left. An alternating current flowing in the coils produces a straight line trace on the screen of the c.r.t.

If a magnetic field is set up in a horizontal direction, the electron beam is deflected in a vertical direction. Two sets of fixed coils mounted at right angles to each other around the neck of the c.r.t. provide control of the beam in both horizontal and vertical directions. By adjustment of the currents flowing in these coils, the electron beam may be directed to any point on the screen of the c.r.t. This method of deflection is commonly used for type *B* presentations.

An alternative method of directing the beam involves the use of only one set of deflection coils mounted in the form of a yoke which may be rotated about the neck of the c.r.t. This method is generally employed for *PPI* presentations. The sweep is made to begin at the centre of the tube and to progress radially to the outside at a linear rate. This requires a current in the deflection coil which begins at zero and increases linearly during the sweep. The yoke is rotated in synchronism with the radar aerial. Hence the direction of the sweep on the c.r.t. indicates the direction in which the aerial is pointing at any instant. Fig. 16-25 illustrates the position of the deflection coils, the linear rise of current and the direction of the sweep when the aerial is pointing eastward. The following paragraphs in this section deal with the circuits used to provide the linearly rising current required for this type of sweep.

The deflection coils contain inductance and resistance. In order to produce a linear rise of current through a resistance a sawtooth voltage must be applied across the resistance, Fig. 16-26 (a). To produce a linear rise of current through an inductance, that is, a constant rate of change of current through it, a square wave of voltage must be applied across the inductance, Fig. 16-26 (b).

If the resistance of the coil and its

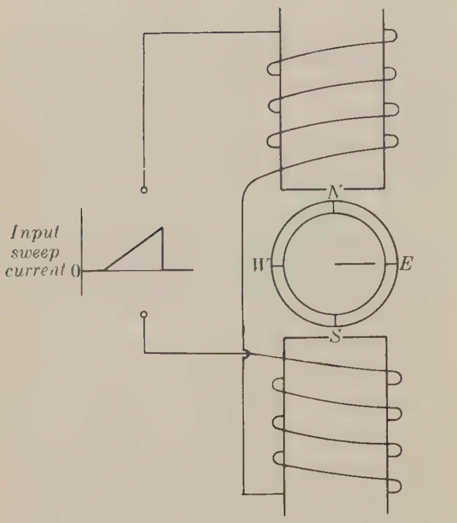


FIG.16-25 Trace produced by sawtooth current in deflection coils.

inductive reactance are comparable the effects of both must be considered. The coil then behaves as a series circuit containing inductance and resistance. The waveform of voltage which must be applied to such a circuit to produce a linearly rising current through it is found by combining the voltage waveforms of Fig. 16-26 (a) and (b). The resulting voltage shown in (c) is called a *trapezoidal waveform*. A sweep generator which produces such a voltage is called a *trapezoidal generator*.

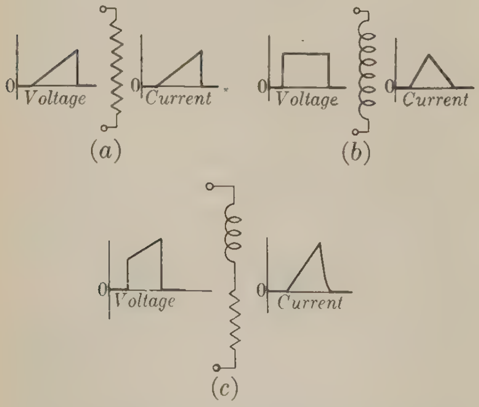


FIG.16-26 Voltage and current waveforms (a) in a pure resistance (b) in a pure inductance (c) in a coil containing resistance and inductance.

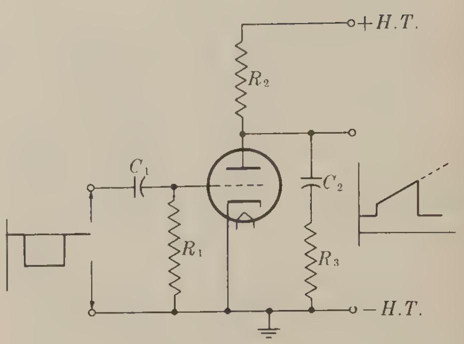


FIG.16-27 Trapezoidal voltage generator circuit.

A. Trapezoidal Generator

A circuit used to generate a trapezoidal waveform of voltage is shown in Fig. 16-27. It is similar to the conventional triode cut-off sweep generator circuit (Fig. 16-8) for generating a sawtooth waveform of voltage with the addition of R_3 and with R_1 connected between grid and ground instead of grid and +H.T. R_3 is connected between the lower plate of the sweep capacitor C_2 and ground. It produces the initial sudden rise of output voltage called the *jump voltage* at the beginning of the sweep.

The valve normally conducts heavily as there is no cathode bias and the grid is returned to ground by R_1. A large p.d. is developed across R_2 which maintains a low anode potential. C_2 charges up to the anode potential and no voltage exists across R_2 during the period between sweeps. When the sweep is to begin, a large negative square wave voltage is applied to the grid of V_1 which cuts off the current through this valve. Since the p.d. across C_2 cannot increase instantaneously, a redistribution of the supply voltage takes place across R_2, C_2 and R_3 in series. The initial current which flows to charge C_2 produces a sudden initial rise in p.d. across R_3. Hence the output voltage is suddenly increased by this p.d. (Fig.

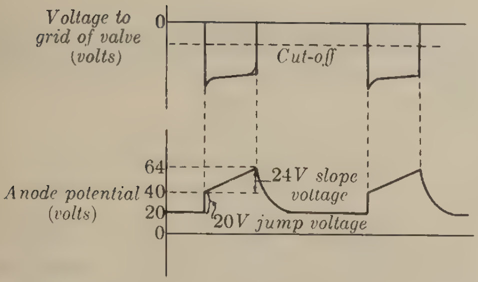


FIG.16-28 Voltage waveforms of a trapezoidal voltage generator.

16-28). Thereafter the sweep capacitor charges exponentially toward the supply voltage at a rate determined by the time constant of R_2, R_3 and C_2. The increase in p.d. across C_2 is limited to a small fraction of the supply voltage so that the sweep is practically linear. At the end of the sweep the grid potential of the valve is suddenly returned above cut-off. The valve conducts again and C_2 discharges to its original p.d. through R_3 and the valve.

The production of the trapezoidal voltage is more readily understood by consideration of a typical example and the waveforms of Fig. 16-28. In the circuit of Fig. 16-27,

$$R_2 = 130{,}000\Omega$$
$$R_3 = 10{,}000\Omega$$
$$C_2 = .005\mu F$$
$$\text{H.T.} = 300 \text{ volts.}$$

Initial anode potential and p.d. across $C_2 = 20$ volts. At the beginning of the sweep, 300 volts is applied across R_2, C_2 and R_3 in series with a p.d. of 20 volts already existing across C_2. The charging voltage is then 280 volts and the resistance in series with C_2 is 140,000Ω. Therefore the initial charging current found from Ohm's Law is $\frac{280}{140{,}000} = .002A$ or $2mA$. The initial rise in voltage across R_3, also found from Ohm's Law, is $.002\times10{,}000 = 20$ volts. Hence the anode potential rises suddenly from 20 volts to 40 volts. The charge gained by C_2 during the sweep is found from the formula $Q = It$ where the current is assumed to be constant. If the sweep continues for 60 microseconds, the charge gained by C_2 is $.002\times60\times10^{-6}$ or $.12\times10^{-6}$ coulombs. The increase in p.d. across C_2 is then given by the formula $Q = CV$. Substituting,

$$.12\times10^{-6} = .005\times10^{-6}\times V$$

and solving, $V = 24$ volts.
Therefore the anode potential rises from 40 volts to 64 volts during the sweep.

At the end of the sweep the valve conducts and its resistance falls to a low value. This value as found from the data given for the initial conditions is about 9300Ω. C_2 discharges through the valve and R_3 which have a total resistance of 19,300Ω. In parallel with the valve is the resistance of R_2 and the power supply. Since the resistance of R_2 is very large its effect in discharging C_2 may be neglected. The time constant of C_2R_3 and the valve is then $19{,}300\times.005\times10^{-6}$ or about 96 microseconds. The time required for the anode potential to fall to its normal value is therefore considerably longer than the sweep time in this instance. This is characteristic of trapezoidal generators but is no disadvantage in radar sets as the sweep time is always followed by a comparatively long waiting period in the complete cycle of operation.

B. Sweep Power Amplifier

The sweep power amplifier V_1 is usually a triode or power valve which supplies the changing current through the sweep deflection coil (Fig. 16-29). This coil is mounted as a rotatable yoke around the neck of the cathode-ray tube. Current is supplied to it by way of slip rings. The valve is biased to cut-off by the cathode bias battery so that no current flows through the deflection coil between sweeps. When the sweep is to begin, the trapezoidal wave-

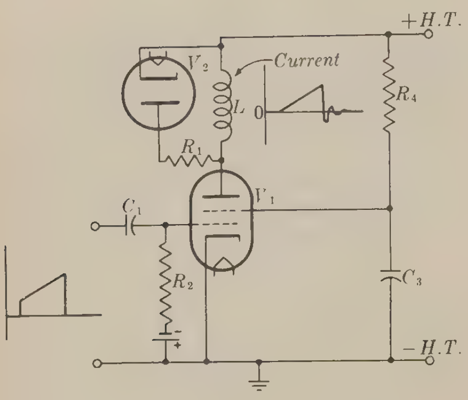


FIG.16-29 Sweep power amplifier circuit.

form of voltage is applied to the grid of the valve. The bias is overcome and the valve is made to conduct current. The current through the deflection coil rises linearly producing a linear sweep from the centre of the screen to the outside. At the end of the sweep, the grid voltage again falls below cut-off. The current through the deflection coil is therefore cut off. The magnetic field about the c.r.t. collapses and the sweep returns to the centre of the tube.

The collapsing magnetic field at the end of the sweep induces a back e.m.f. which tends to maintain the current flow in the deflection coil. This causes oscillation in the coil at a frequency determined by its inductance and self-capacitance. This oscillation must be damped out before the succeeding sweep commences so that linearity of the sweep may be maintained. Damping is provided by the resistor R_1 which is connected in series with diode V_2 across the deflection coil. The diode does not conduct during the sweep since its cathode is more positive than its anode, and so R_1 provides no damping effect across the coil. At the end of the sweep, however, the oscillation causes the anode end of the coil to rise to a higher potential than the supply voltage. The anode of the diode is then positive to its cathode and the valve conducts. This current flowing through R_1 provides the damping which quickly reduces the oscillation to zero.

C. Multi-Range Sweep-Current Generator

A sweep-current generator and amplifier circuit which provides deflection currents for three different ranges on the *PPI* tube is shown in Fig. 16-30. This circuit provides automatic compensation for changes in operating conditions of the valves brought about by the range switching. V_1 is the trapezoidal generator valve. The sweep selector switch S selects the proper value of anode resistance and sweep capacitance required to provide sweep voltages of the required magnitude. The duration of the sweep is controlled by the duration of the negative square wave applied to the grid of V_1. The waveforms of V_1 grid and anode voltages for the three ranges are shown in Fig. 16-31 (a) and (b) respectively. If, for example, the required ranges are 10,000 yards, 20,000 yards and 50,000 yards, the durations of the negative square waves to the grid of V_1 must be 61, 122 and 305 microseconds respectively. The sweep capacitors are then chosen so that the linear rise in voltage during the sweep has the same amplitude for each range. This ensures that the trace is of the same physical length on the screen of the c.r.t. for each of the three ranges. The variable resistors R_1, R_2 and R_3 are chosen for this condition.

The amplitudes of the jump voltage at the beginning of the sweep are different for each of the ranges. For the short range, the current through the deflection coil L must rise from zero to its maximum value in less time than for the other ranges. Consequently a larger amplitude of jump voltage is required because of the large inductance effect of the coil.

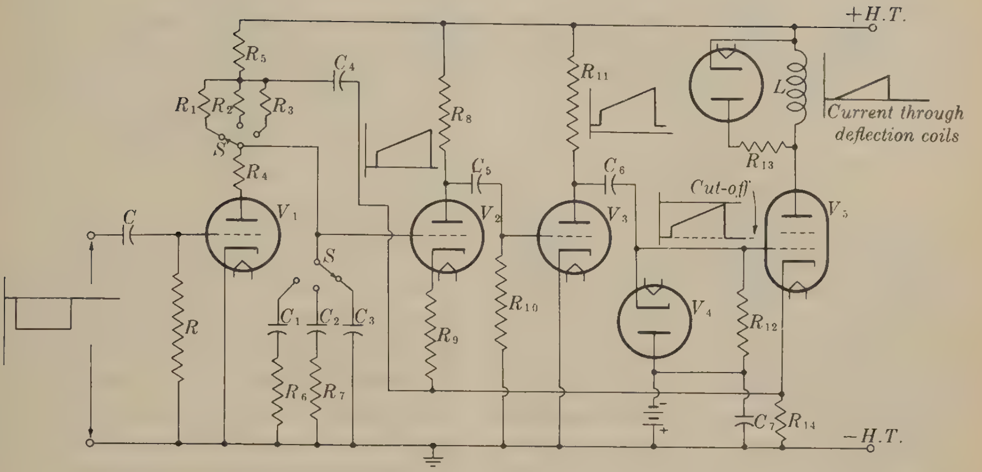


FIG. 16-30 Multi-range sweep current generator.

On the longest range, the current rises so slowly through the coil that the inductive effect is almost negligible and the jump voltage is no longer required. Since the coil is effectively resistive the current through the coil is in phase with the voltage applied across the coil.

V_2 and V_3 are voltage amplifiers which amplify and invert the sweep voltage twice before applying it to the grid of the power amplifier. The output voltage waveforms of V_2 and V_3 for the three ranges are shown in Fig. 16-31 (c) and (d). These waveforms are not drawn to the same scale and do not show the increase in amplitude due to amplification. Degeneration is introduced in V_2 by omitting the cathode by-pass capacitor. Bias is provided by the cathode resistor to enable this valve to handle the positive-going sweep voltage. No bias is required for V_3 since its signal is a negative-going voltage.

The waveforms of Fig. 16-31 indicate the function of the diode V_4. This valve is a clamping valve in parallel with the grid-leak resistor of V_5. In the absence of V_4, the voltage at the grid of V_5 would be a voltage varying about the bias level because of the blocking action of the coupling capacitor, Fig. 16-31 (e). Due to the unsymmetrical nature of the waveforms, the grid would not operate over the same range of positive and

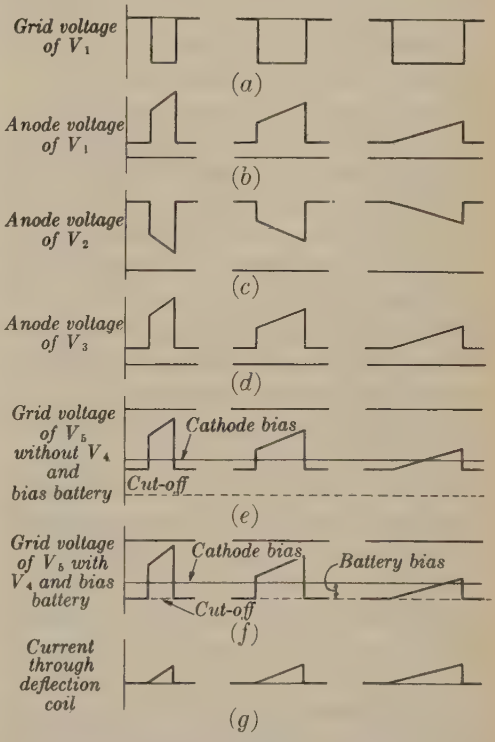


FIG. 16-31 Voltage and current waveforms of multi-range sweep generator.

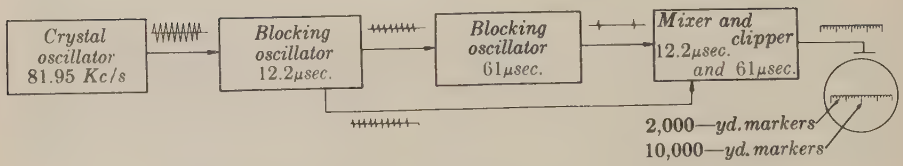

Crystal oscillator 81.95 Kc/s
Blocking oscillator 12.2μsec.
Blocking oscillator 61μsec.
Mixer and clipper 12.2μsec. and 61μsec.
2,000—yd. markers
10,000—yd. markers

FIG. 16-32 Block diagram of sweep calibration circuit.

negative potentials for each of the three sweeps. Consequently the sweep could not be made to begin at the centre of the tube and extend to the outside on each of the three ranges. This defect is overcome by placing the clamping diode V_4 in parallel with the grid-leak resistor. It ensures that the coupling capacitor always becomes discharged to the same voltage during the waiting period between sweeps. The sweep voltage is therefore clamped to operate on the positive side of the bias potential only. By proper choice of the bias voltage, the sweep current may be made to begin at zero at the beginning of each sweep. This ensures that the centre of the c.r.t. screen is the point of zero range.

The linearity of the sweep voltage waveform is improved by two feedback circuits from the cathode of V_5. The first of these is to the cathode of V_2 from the top of the un-by-passed cathode resistor R_{14} which provides degeneration. Suppose the sweep voltage to the grid of V_5 contains a non-linear portion due to distortion introduced by V_2 and V_3. A fall in voltage at the grid of V_5 produced a fall in voltage at the cathode of V_5 and hence at the cathode of V_2. This is equivalent to a rise in voltage at the grid of V_2. The result is a fall in voltage at the anode of V_2 and a rise in voltage at the anode of V_3. The fall in grid voltage of V_5 is therefore corrected by an amplified rise in voltage through the feedback circuit.

The second feedback line is from the cathode of V_5 to the upper end of the sweep resistors in the anode circuit of V_1. This circuit operates in a manner similar to that of Fig. 16-13. As the sweep voltage rises at the grid of V_2, the cathode potential of V_5 also rises. This rising voltage is fed back through C_4 to the anode circuit of the trapezoidal generator. It effectively increases the supply voltage to the sweep circuit as the sweep progresses. This tends to maintain a linear rise of voltage across the sweep capacitor.

16.5 Calibration of sweeps.

A. Calibration of a Type A Scan

Reference has already been made to the calibration of *A* scan sweeps by means of a shocked oscillator (Fig. 15-34). A more complete circuit is suggested by the block diagram of Fig. 16-32. It provides a series of downward pips along the trace which are spaced 12.2 microseconds apart in time and which therefore represent 2000 yard ranges between target and radar set. This relationship is evident when it is recalled that radar pulses and echoes travel at the speed of light, 162,000 nautical miles per second. The time required for a pulse to travel one mile is therefore $\frac{1}{162{,}000}$ of one second or 6.2 microseconds. Since 1 nautical mile is 6080 feet, the time required for a pulse to travel 2000 yards is $\frac{2000}{162{,}000 \times \frac{6080}{3}}$ or 6.1 microseconds. The time required for a pulse to travel to a target 2000 yards away and for the echo to return is therefore 12.2 microseconds. Similarly the time taken by a pulse in travelling to and returning from a target 10,000 yards distant is 61 microseconds. A crystal oscillator is used as the timing device because of its excellent frequency stability. It operates at a frequency of 81.95 Kc/s and therefore produces one cycle every 12.2 microseconds. The output sine wave of voltage is applied to the grid of a biased-off blocking oscillator. This circuit produces a single oscillation of short duration for every cycle of input. These pulses are supplied to a second biased-off blocking oscillator which has a somewhat longer recovery time. This circuit produces one oscillation for every five input pulses it receives. These pulses, occuring at 61 microsecond intervals are applied, together with the 12.2 microsecond pulses, to the grid of a clipper and mixer valve. This circuit produces a series of output negative pulses which are applied to the top *Y*-plate of the c.r.t. They repel the electron beam as it sweeps across the screen of the tube, producing downward pips along the scan. The 10,000-yard markers are usually made larger than the 2000-yard markers to make them more easily read.

The calibration markers must be synchronized with the start of the sweep to prevent them from drifting along the trace. This may be done by turning the oscillator on at the beginning of each sweep and turning it off at the end of the sweep. A more satisfactory method with this type of circuit is to permit the crystal oscillator to operate continuously and to use it to control the start of the sweep and the pulsing of the radar transmitter. This is done by further subdividing the crystal oscillator frequency by a blocking oscillator or multivibrator. For example, if the crystal frequency, 81.95 Kc/s, is sub-divided by a factor of 10 in each of two successive stages, a well stabilized frequency of 819.5 cycles per second is obtained. This frequency is used as the pulse repetition frequency of the radar set. The sweep is started and the transmitter is pulsed once during each cycle, that is every 1220 microseconds. The marker frequency is then an exact multiple of the repetition frequency and the markers do not drift along the trace.

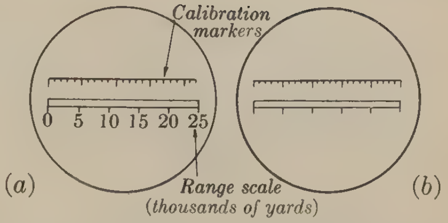


FIG.16-33 Calibration of type A presentation (a) Sweep speed too fast for coincidence between calibration markers and fixed scale marks (b) Sweep speed correctly adjusted.

Calibration markers may be permitted to remain on the screen during normal operation of the set, or may be used merely to check the speed of the sweep against a fixed range scale on the front of the c.r.t. This function is illustrated in Fig. 16-33 (a) and (b). Fig. 16-33 (a) shows the appearance of the c.r.t. trace and range scale when the sweep speed is too fast to bring the calibration markers into coincidence with the fixed scale. The calibration markers are displaced to the right. A slight increase in the resistance or capacitance in the sweep generator is required to increase the time of the sweep charging circuit. This results in a decrease in the rate at which the electron beam is swept across the screen, causing the calibration markers to appear more closely crowded together. The adjustment of the sweep generator is made with the c.r.t. and calibration circuits in operation. In this way the calibration markers are brought into coincidence with the corresponding marks on the range scale as shown in Fig.16-33 (b).

It is not the usual practice to apply the calibration markers to the deflection plates continuously as small echo pulses

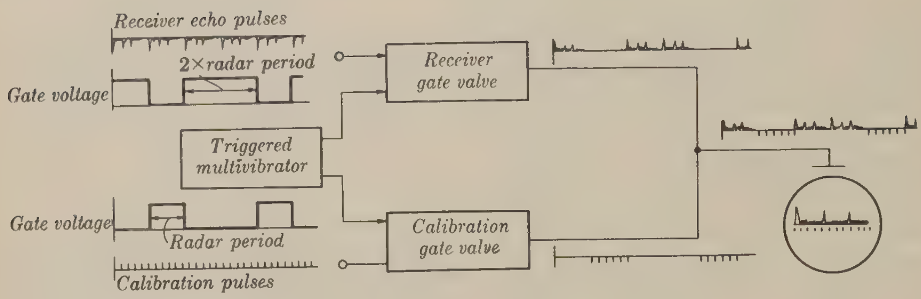


FIG.16-34 Block diagram of a gating circuit.

may be obscured by the calibration markers. One method is to turn on the calibration markers for a short period only while ranging on a target. Alternatively, the calibration markers may be switched on electronically during every third or fourth sweep and the radar echo pulses during the other sweeps. This is accomplished by a gating circuit which operates one of the amplifiers in each of the receiver and calibration circuits.

B. Gating Circuits

A typical gating circuit is shown in block diagram form in Fig. 16-34. An unbalanced, biased-off multivibrator is triggered by synchronizing pulses at the pulse repetition rate of the radar set. When this circuit receives a synchronized pulse, it goes through one cycle of oscillation. The long half cycle has a duration equal to twice that of the radar period, that is, twice the time between radar pulses. The short half cycle has a duration equal to the radar period. The multivibrator therefore operates at a frequency equal to one-third of the pulse repetition frequency of the set. Out-of-phase voltages of rectangular shape are obtained from opposite anodes of the multivibrator. These are applied to the suppressor grids of a receiver gate valve and a calibration gate valve respectively. The gate voltage permits each valve to amplify in its normal manner during the positive half of the gating cycle and cuts off the flow of anode current during the negative half of the cycle. In effect then, the two valves are turned on alternately. The receiver gate valve permits the positive echo pulses to pass to the upper Y-plate of the c.r.t. during two successive sweeps and cuts off this input during the next sweep. The calibration gate valve permits the positive calibration pulses to pass to the upper Y-plate during one sweep and blocks this path during the following two sweeps.

The circuit of the calibration gate valve is shown in Fig. 16-35. The valve is biased to cut-off by a positive potential on the cathode from the voltage divider R_3, R_4 and R_5 across the H.T. supply. Positive calibration pulses, applied to the control grid of the valve, overcome the cathode bias and permit corresponding pulses of current to flow through the valve. Each of these develops a p.d. across the anode load resistor R_2 causing a negative calibration pulse at the anode. In addition to the calibration pulses applied to the control grid, the gate voltage is applied to the suppressor grid. It maintains the suppressor at ground potential for normal operation during the short half cycle. During the long half cycle the suppressor is driven sufficiently negative to prevent the flow of anode current. The anode potential remains at H.T. during this period and so the signal on the control grid does not appear in the output.

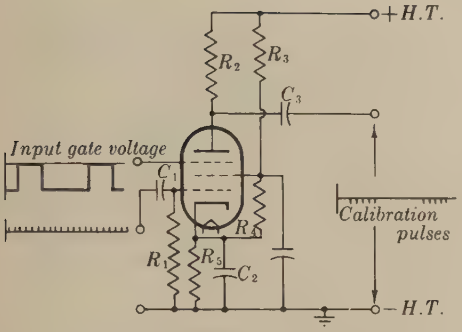


FIG.16-35 Gate valve circuit.

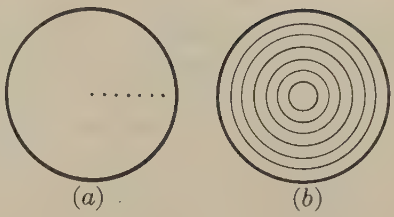


FIG.16-36 Calibration of P.P.I. presentation (a) Calibration dots with scan at rest (b) Calibration rings with scan rotating.

C. *Calibration of PPI Scan*

The calibration pulses developed for the *A* type scan may also be applied to the *PPI* tube. They produce intensity modulation of the electron beam in a manner similar to the echo pulses. The *PPI* tube is operated with the intensity grid potential just below cut-off. Positive calibration pulses are applied to the intensity grid, or negative pulses to the cathode. In either case, the bias is overcome and the electron stream strikes the screen at intervals while the sweep takes place. The effect produced on the c.r.t. during a single sweep is shown in Fig. 16-36 (a). If the sweep is not rotated, this pattern is repeated at the sweep repetition rate and a series of bright dots is observed. If the sweep is rotated a series of concentric rings is seen as shown in Fig. 16-36 (b). Each ring gives a measure of the range of the target from the radar set. The farther an echo pulse appears from the centre of the screen, the greater the range to the target. Calibration rings are not normally applied continually as they may obscure small echoes. They are turned on only during calibration and ranging procedures.

16.6 Accurate ranging methods. The *A* scan and *PPI* scan as previously described are termed *general viewing scans.* They are used in surface warning and aircraft warning equipments in which accurate determination of range to the target is not of primary importance. Fire control equipments require special circuits for determining range with the utmost accuracy and speed. This involves a precise measurement of the time interval between the beginning of the transmitter pulse and the beginning of the echo pulse. A method widely used for accurate ranging employs the movable *range step* illustrated in Fig. 16-37. The range step is a sharp fall in the trace as shown in (a). It may be made to occur at any position along the trace. Its position is controlled by a handwheel on the front panel of the indicator unit. This handwheel operates a *delay multivibrator* which begins its cycle at a predetermined time before the beginning of the sweep and produces a positive square

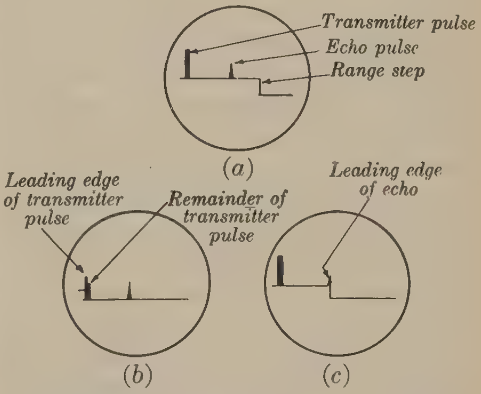


FIG.16-37 Range step on type A presentation (a) General appearance of range step (b) Position of range step with range dial at zero (c) Ranging on a target.

pulse, which is applied to the upper Y-plate. The trailing or negative-going edge of the pulse occurs at any desired time during the sweep and produces the range step. The time between the start or leading edge of the positive square pulse and the trailing edge is called the *delay* of the multivibrator. As the handwheel is turned clockwise, the delay of the multivibrator cycle is increased and the step is observed to move to the right. The handwheel is also geared to a range dial which indicates the distance between the transmitter pulse and the range step in yards or miles. When the range step is moved to the extreme left, Fig. 16-37 (b), so that it is opposite the leading edge of the transmitter pulse, the range dial indicates zero range. When the range step is moved to the right hand side of the trace, the range dial indicates maximum range. Ranging on a target then consists of moving the range step until it is opposite the leading edge of the echo pulse, Fig. 16-37 (c), and reading the range from the range dial.

A. A Basic Delay Multivibrator

A basic type of delay multivibrator which may be used to produce the range step is shown in Fig. 16-38. The circuit is a biased-off, cathode-coupled multivibrator. V_2 is the normally *on* valve as its grid is returned to H.T. through R_5. The grid potential is limited to about 1 volt positive to the cathode, which is considerably positive to ground because of the flow of anode and grid currents through R_7. V_1 does not conduct in the normal state. Its grid is slightly positive to ground because of the voltage divider R_1, R_2 across the H.T. supply. However, the cathode is considerably more positive than the grid so that the valve is cut off. In the stable condition of the circuit, the anode potential of V_1 is at H.T. while that of V_2 is considerably lower.

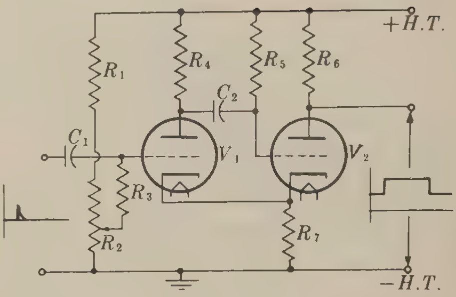


FIG.16-38 Range-step generator circuit.

The multivibrator cycle is started by applying a sharp positive pulse to the grid of V_1 via C_1. This causes V_1 to conduct and its anode potential to fall. C_2 discharges through R_5 developing a negative voltage at the grid of V_2 which cuts off the current through this valve. The time during which V_2 is cut off is the delay and is determined by the time constant of C_2R_5 and also by the cathode potential of V_2. This potential is determined by the amount of current flowing through V_1 which in turn is controlled by the setting of R_2. If the movable tap of R_2 is moved upward, the grid potential is made more positive and an increased current flows through the valve V_1. This current flowing through R_7 causes an increased positive potential at the cathode of V_2. A rise in cathode potential is equivalent to a drop in anode-to-cathode voltage and therefore decreases the negative cut-off potential. Therefore the time taken for the grid of V_2 to return to cut-off potential is increased and the multivibrator delay is extended. During the period that V_2 is cut off, there is no current flow through R_6 and no p.d. developed across it. The anode potential is therefore at the H.T. value. When the valve conducts again the anode potential falls. This fall in the anode voltage is applied to the top Y-plate of the c.r.t. to produce the range step. The waveforms of grid and anode potentials of V_2 and the effect produced on the

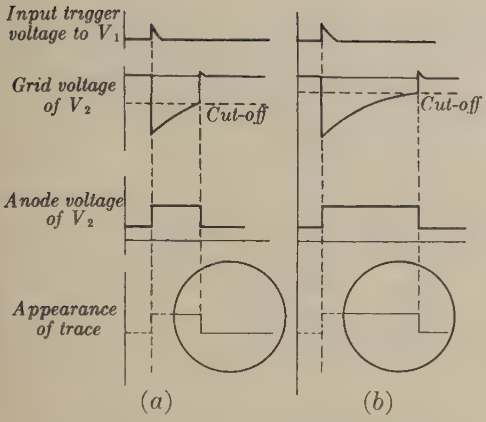


FIG.16-39 Effect of cathode potential of V_2 on position of range step (a) High cathode potential produces short delay (b) Low cathode potential produces long delay.

indicator trace are shown in Fig. 16-39 (a). The effect of increasing the grid potential of V_1 is shown in Fig. 16-39 (b). The duration of the positive square pulse at the V_2 anode is increased and the range step is moved farther to the right along the trace. The leading edge of the multivibrator output does not show on the c.r.t. This is because it must be possible to place the step at the start of the sweep for zero range. Therefore the multivibrator cycle must start some time before the sweep starts since the duration of the positive output square pulse at the V_2 anode has a minimum value. The timing circuit of the radar set starts the multivibrator cycle and delays the transmitter pulse and the beginning of the trace for this minimum duration. Resistor R_2 controls the duration of the positive square pulse from V_2 and hence the position of the range step on the trace. It is operated by the handwheel which also operates the range dial.

While the basic operation of the range step generator is simple, the problem of tracking the mechanical range dial to the electrical delay circuit presents some difficulty. The grid voltage curve is not a linear one, but an exponential one. An increase in the cathode potential level of one volt does not produce the same change in duration of the positive square pulse output from V_2 over the entire range of operation. Thus a linear dial may be properly calibrated at the minimum and maximum ranges and be considerably in error for intermediate readings. More elaborate circuits than the basic type have been designed to make the multivibrator delay a linear function of the resistance controlling that delay. The following circuit is an example of a *linear-delay multivibrator.*

B. *Phantastron*

A pentagrid valve of the type used as a frequency converter in radio receivers is shown in the *phantastron* delay circuit of Fig. 16-40. The operation of the circuit is similar to that of the cathode-coupled multivibrator just described. The cathode, grid 3 and anode perform the functions of the *off* valve during the stable state while the cathode, grid 1 and grids 2 and 4 constitute the *on* valve. Typical waveforms and values of the voltages at each of the valve electrodes are shown in Fig. 16-41. In the steady state, all the current flowing from the cathode passes to the first three grids in the valve. This current amounts to $4mA$ flowing through the cathode resistor, which sets the cathode potential at 40 volts. Grid 1 draws only $.06mA$ of current, which flows through the large grid-leak resistor R_3, setting the grid potential also at 40 volts. The screen grid potential, that is, the potential of grids 2 and 4 is determined partly by the voltage divider R_4, R_7 and partly by the screen current flowing through R_4. The combined effect produces a voltage of 69 volts at the screen. The potential of grid 3 is determined by the voltage divider R_1, R_5 and is 27 volts. This grid is therefore 13 volts negative to the

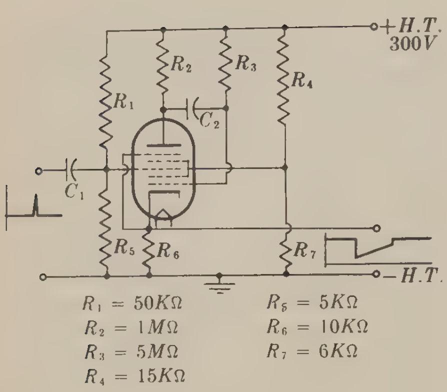


FIG.16-40 Phantastron delay circuit.

cathode, which is sufficient to cut off the flow of current from cathode to anode. Consequently no current flows through R_2 and so the anode potential is at H.T. The circuit continues in this state until triggered by a positive pulse to grid 3. A sequence of changes comprising three phases then takes place. The first phase is a multivibrator action which causes a sudden change in potential at each of the electrodes. This is followed by a relaxation period during which the voltages change linearly with time until critical values are reached. At this point a third set of changes takes place suddenly and the circuit resumes its original steady state conditions. A detailed description of each phase follows.

The positive trigger pulse applied to grid 3 must raise the potential of this grid about three volts to start the switching action. This overcomes the cut-off voltage and permits current to flow to the anode. The anode potential drops sharply and C_2 discharges slightly through R_3, causing a fall in the potential of grid 1. This reduces the current through the valve, so that the cathode potential decreases. A fall in the cathode potential has the same effect as a rise in grid 3 potential. Therefore the anode current, which was begun by the trigger pulse, is now increased by feedback through C_2.

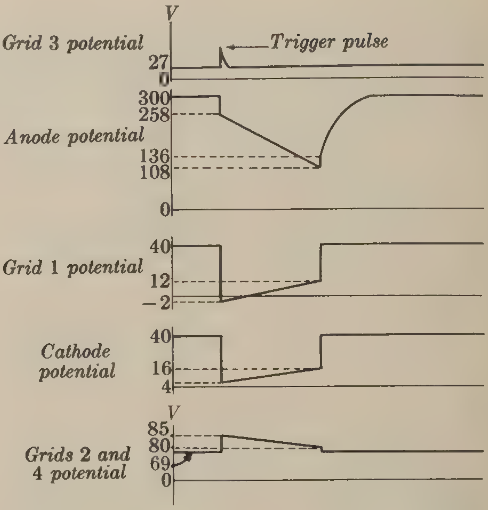


FIG.16-41 Voltage waveforms of phantastron delay circuit.

The process is cumulative to the point where a further reduction in grid 3-to-cathode voltage is no longer effective in increasing the anode current. The cathode potential falls abruptly from 40V to 4V, the grid 1 potential from 40V to −2V and the anode potential from 300V to 258V. The reduction of grid voltage reduces the screen current causing a rise in screen potential from 69 to 85 volts.

The relaxation or *run-down period* now begins during which the potential of grid 1 tends to return to its normal value as in a multivibrator. As the grid potential rises, however, the anode current continues to increase and the anode potential continues to fall. The combined effect of rising grid potential and falling anode potential causes the grid voltage to depart from the exponential curve and to follow a linear rise as shown by the waveforms of Fig. 16-41. The precise operation of C_2 is the same as that in the Miller time-base circuit of section 16.2F. As the grid 1 potential rises, the cathode current increases and the cathode potential rises. This is equivalent to a fall in potential of grid 3 which eventually causes the anode current to stop increasing and start decreasing. A point is therefore reached where a further increase in anode current due to the rise of potential on grid 1 is prevented by the rise in cathode potential which it produces. When this point is reached, a second switching action takes place. The decrease in anode current causes the anode potential to rise and C_2 commences to charge again. The potential of grid 1 rises sharply, permitting the normal cathode current to flow. The anode current is again cut off and the screen current increases to its normal value.

The duration of the phantastron output pulse taken from the cathode is determined by the time taken for the anode potential to fall to its critical value, which is 108 V in this example. If the circuit elements remain constant, the rate of fall of anode potential also remains constant. The duration of the cycle may then be reduced by using a smaller initial potential on the anode of the valve. By connecting the anode to a potentiometer across the H.T. supply, the duration of the phantastron pulse may be made to vary linearly with the setting of this control. Such an arrangement is shown in Fig. 16-42. The circuit includes a diode valve V_1 for isolating the anode of V_2 from the potentiometer R_5 during the phantastron pulse, and a triode V_3 connected in a cathode follower circuit between the anode and grid 1 of the phantastron.

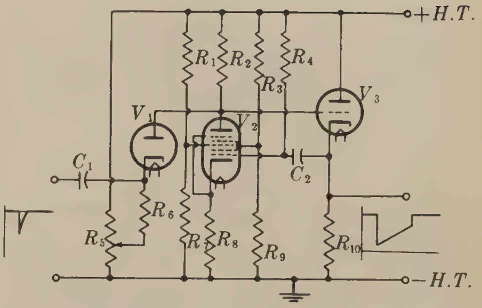


FIG. 16-42 Phantastron sweep generator with cathode follower to improve linearity.

In the steady condition of this circuit, the anode potential is fixed by the setting of R_5. R_5 and R_6, in series with R_2, are small in comparison with R_2 so that a wide range of anode potentials may be obtained. No anode current flows in V_2 during the quiescent period and the other potentials are exactly as described for the basic phantastron circuit. A negative trigger pulse is applied via C_1 to the cathode of V_1. This pulse is applied to V_1 in series with the anode resistor of V_2. Since R_2 is a large resistor, most of the voltage change appears across R_2. The negative pulse is therefore developed at the anode of V_2 and at the grid and cathode of V_3. It is applied in the same polarity to grid 1 of V_2 via C_2. This trigger pulse reduces the cathode current and cathode potential momentarily and thereby starts the flow of anode current. A further decrease in anode potential takes place, which is fed via V_3 and C_2 to the grid 1 of the phantastron valve. The switching action therefore continues, independent of the triggering pulse.

As the anode potential falls, the anode of V_1 becomes negative to its cathode and so this valve stops conducting. It therefore acts as an open circuit between the anode of V_2 and the potentiometer circuit preventing the latter from loading the anode circuit. The cathode follower serves mainly to recharge C_2 after the cycle has been completed. This capacitor which was required to charge through the high resistance of the anode load resistor in the basic circuit is now able to charge through the low cathode-anode resistance of V_3. This permits the anode potential of the phantastron to resume its normal value in a short time, eliminating the trailing edge of the waveform shown in Fig. 16-41.

The phantastron circuit may also be used to develop the range step for

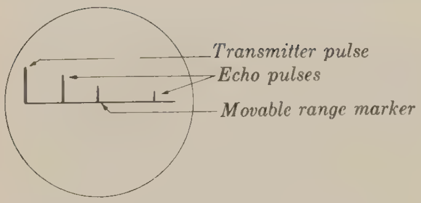


FIG. 16-43 Range marker on type A presentation.

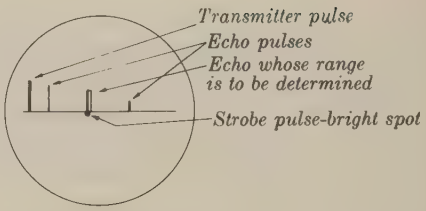


FIG. 16-44 Strobe pulse on type A presentation.

accurate ranging on radar echoes. The output negative pulse from the V_3 cathode must be squared by additional circuits before application to the c.r.t. A potentiometer with a linear relationship between angle of rotation and resistance value is used as the range potentiometer R_5. It is operated by the handwheel for shifting the range step. The range dial which is geared to the handwheel is linearly calibrated to track with the range step throughout the entire trace on the c.r.t.

C. *Variations of Range Step*

The principle of the range step operated by a handwheel and aligned with a range dial may be applied in a number of ways. These include the *range marker*, *strobe pulse*, *range notch* and *range pedestal*. A brief reference to each of these types follows, together with the appearance of the c.r.t. trace in each case.

1. RANGE MARKER.

The trailing edge of the square pulse which appears as the range step may be differentiated to form a sharp pulse, inverted by an amplifier and applied to the top Y-plate to produce an upward pip on the trace. It is similar in appearance to an echo pulse and is shown in Fig. 16-43. The method of ranging on a target is similar to that using a range step. The marker is shifted along the trace until its leading edge coincides with the leading edge of the echo and the range read from the range dial.

2. STROBE PULSE.

If the positive range marker pulse referred to above is applied to the intensity grid instead of the upper Y-plate of the c.r.t., the electron beam is increased in intensity and a bright spot appears on the trace. Such a pulse is called a *strobe pulse*. The strobe spot may be moved to any position along the trace as in the case of the range marker. It is adjusted to illuminate the leading edge of the echo (Fig. 16-44) and the range is read from the range dial.

3. RANGE NOTCH AND RANGE PEDESTAL.

The trailing edge of the square pulse from the range step generator may also be used to trigger a second square wave generator. The output of the latter circuit is then applied as a negative square wave to the upper Y-plate to produce a *range notch* as shown in Fig. 16-45 (a). Alternatively, a positive square wave may be applied to the upper Y-plate to produce a *pedestal* as in Fig. 16-45 (b). In either case, the notch or pedestal is shifted by the delay circuit until its leading edge coincides with the leading edge of the echo pulse. The range is then given by the range dial.

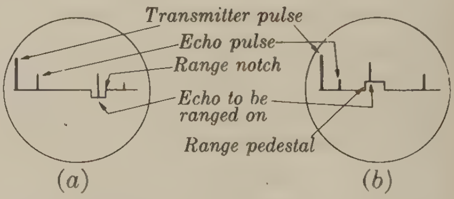


FIG. 16-45 Range notch and range pedestal on type A presentation.

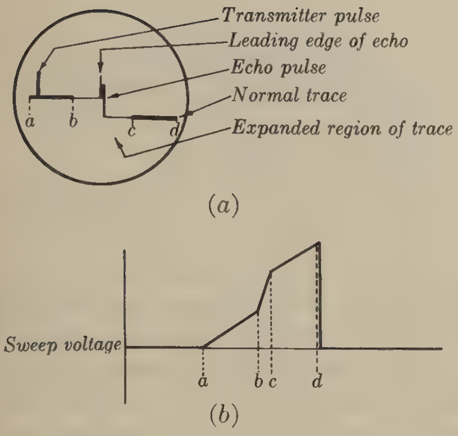


FIG.16-46 Expanded sweep on type A presentation (a) Appearance of trace on c.r.t. (b) Waveform of sweep voltage applied to right X-plate.

D. *Expanded Sweeps*

The problem of aligning the leading edge of a range step, range marker, notch or pedestal with the leading edge of an echo is one which may limit the accuracy of the radar set to a serious degree. This is especially important in the case of long-range sweeps in which an error of one-tenth of an inch in setting the range step along the trace corresponds to several miles error in range to the target. In order to facilitate a more accurate setting, an *expanded sweep* is sometimes used. Such a sweep in effect magnifies a small area of the trace in the region containing the echo and range step. The effect is shown in Fig. 16-46 (a). The sweep is made to progress at a normal rate from *a* to *b*. In the region from *b* to *c* which includes the range step, the sweep is made to progress more quickly than normally from left to right. This effect is produced by causing the deflecting voltage on the right *X*-plate to rise more steeply during the interval *b* to *c*, Fig. 16-46 (b). The trace and echoes appear somewhat fainter in this region because there are fewer electrons striking the screen per unit of length of scan. In the region from *c* to *d* the sweep is made to proceed at its normal slower rate again so that the trace is as bright as in the interval from *a* to *b*.

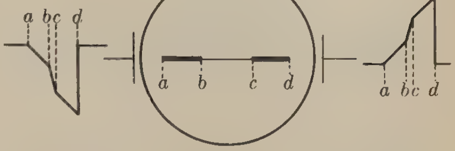


FIG.16-47 Sweep voltages used to expand sweep.

The expanded sweep may be produced by electrostatic or electromagnetic deflection. In either case, the fast sweep is superimposed upon the slow sweep. The complex waveforms of voltage applied to the *X*-plates is shown in Fig. 16-47. A circuit used to produce such a sweep voltage is shown in Fig. 16-48. The circuit is a constant-current-pentode sweep generator similar to that shown previously in Fig. 16-11. The pentode valve V_2 is connected in series with a triode V_1 across the H.T. supply. V_1 is unbiased as its grid is returned to its cathode. Hence only a small part of the supply voltage appears across this valve. The pentode valve is biased in its cathode circuit while its grid is returned

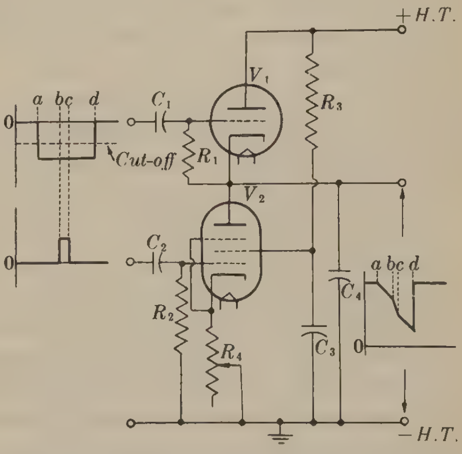


FIG.16-48 Circuit used to produce expanded sweep voltage.

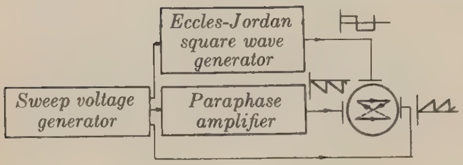


FIG.16-49 Block diagram of double-sweep circuits.

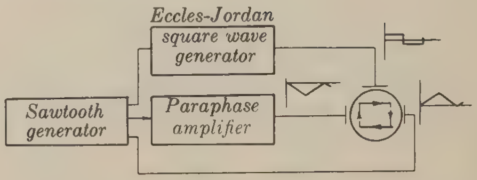


FIG.16-50 Rectangular type of double sweep.

to ground. In the normal state the anode potential of V_2 is high and the sweep capacitor C_4 is charged to this potential. When the sweep is to commence, at time *a* in the diagram, the grid potential of V_1 is made sufficiently negative by a negative square pulse applied via C_1 to cut off the flow of current through this valve. C_4 begins to discharge through the cathode resistor R_4 and V_2. This produces a linear fall of voltage across C_4 indicated by *ab*. At this point, a positive square pulse of voltage is applied to the grid of V_2. This reduces the bias on V_2 and causes it to conduct a larger current during the period *bc*. The sweep capacitor therefore discharges at a faster rate and the sweep voltage falls more steeply. At *c* the grid voltage to V_2 is returned to normal and so the sweep capacitor discharges at its initial rate from *c* to *d*. The voltage across C_4 is applied to the left *X*-plate of the c.r.t. A paraphase amplifier inverts this voltage waveform and applies it to the right *X*-plate.

The timing of the square pulse to the grid of V_2 is controlled by a delay circuit similar to that which develops the range step. It is arranged to begin the pulse at time *b* shortly before the range step is produced and to continue it to time *c* shortly after the range step ends. In this way the range step always appears in the expanded sweep region. The range step may then be accurately aligned with the leading edge of the echo pulse and an accurate determination of range to the target obtained from the range dial.

E. Double Sweeps

For accurate measurement of time intervals, it is desirable to extend the physical length of the sweep. In this way a longer line may be used to represent the same time interval as the single line trace considered previously. A number of methods have been devised for extending the length of the sweep without using a larger c.r.t. One of these is the *double-sweep method* shown in Fig. 16-49. The sweep voltage generator is made to operate at double its normal frequency so that two sweeps are produced in the time formerly taken by a single sweep. The two sweeps are separated from each other by a square wave of voltage applied to the *Y*-plates. The *trace-separation* voltage is obtained from an Eccles-Jordan circuit which receives its input from the sweep generator. At the start of the first sweep, the sweep voltage generator triggers the *E-J* circuit which in turn applies a steady positive potential to the upper *Y*-plate of the c.r.t. During the flyback time, the sweep generator triggers the *E-J* circuit again. This circuit then applies a steady negative potential to the upper *Y*-plate which depresses the scan during the next sweep. The path of the electron sweep in the c.r.t. is indicated by the arrows. This type of sweep is not generally used in radar sets as there is a possibility of echoes being lost during the flyback time between the sweeps. It does, however, find application in electronic navigational-aid equipment.

A variation of the double-sweep method is illustrated by the block diagram of

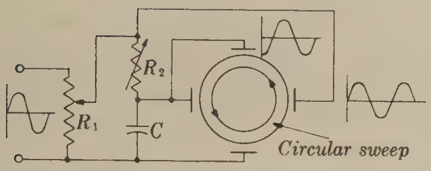


FIG.16-51 Basic operation of circular-sweep generator.

Fig. 16-50. The sawtooth generator produces a linear forward sweep as in the previous circuit, followed by a linear backward sweep of the same duration. This voltage is amplified and inverted and is applied to the X-plates of the c.r.t. The electron beam is swept at a uniform rate from left to right, then from right to left. These two sweeps are separated as before by an Eccles-Jordan circuit which is triggered at the beginning of each forward and backward sweep. As a result, the trace follows the rectangular pattern indicated in the diagram.

F. *Circular, Elliptical and Spiral Sweeps*

The physical length of the trace may also be increased by causing the electron beam to trace out a circular pattern on the screen of the c.r.t. The method of producing such a sweep has already been suggested in Chapter IV under the heading of Lissajous figures. If two sine wave voltages of equal amplitudes and 90° difference in phase are applied to the horizontal- and vertical-deflection plates simultaneously, a circular trace is produced.

The basic circuit used to produce a circular sweep is shown in Fig. 16-51. A sinusoidal waveform of voltage is applied across R_2 and C in series. The same current therefore flows through each of these components. The voltage developed across R_2 is in phase with the current while the voltage across C lags the current by 90°. The voltage across R_2 is used as the horizontal-deflection voltage while that across C is applied to the vertical-deflection plates. The direction of the sweep may be determined by analyzing the instantaneous voltages applied to the X- and Y-plates at 90° intervals during one complete cycle (Fig. 16-52). At time a, the voltage applied to the horizontal-deflection plates is zero while that applied to the vertical-deflection plates is a maximum negative. The spot therefore appears at position a in the c.r.t. A quarter cycle later, at time b, the horizontal-deflection voltage is a maximum positive and the vertical-deflection voltage is zero. Consequently the sweep is at point b in the c.r.t. Another quarter cycle later, the horizontal-deflection voltage is again zero while the vertical-deflection voltage is a maximum positive. The spot is therefore at position c in the c.r.t. After another quarter cycle has passed the sweep has advanced to position d. The sweep therefore rotates counter clockwise in this case. The direction of rotation may be reversed by interchanging the horizontal-and vertical-deflection voltages.

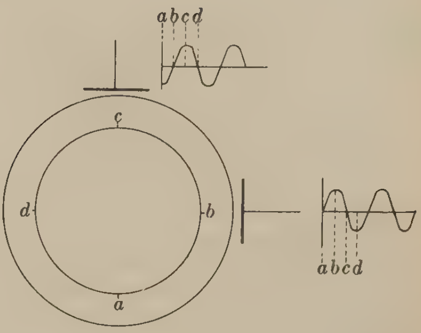


FIG.16-52 Production of circular sweep from sinusoidal voltages 90° out of phase.

If the amplitude of the horizontal-deflection voltage is greater than that of the vertical-deflection voltage, the trace becomes an ellipse with its horizontal axis greater than its vertical axis.. If the vertical-deflection voltage is the greater one, the vertical axis is greater than the horizontal axis. An elliptical sweep may

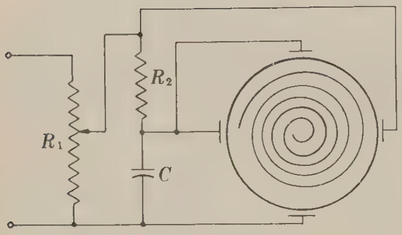


FIG. 16-53 Basic spiral-sweep circuit.

also be produced by applying deflection voltages of equal amplitudes and a phase difference other than 90°.

If the amplitudes of two equal deflection voltages, 90° out of phase, are both increased, a circular trace of larger diameter is produced. This may be accomplished with the circuit of Fig. 16-51 by moving the variable tap of R_1 upward. A spiral trace may be obtained by increasing the amplitudes of the deflection voltages gradually as the sweep progresses. Such a trace is shown in Fig. 16-53. The five loops of the spiral are produced by moving the variable tap of R_1 from its minimum to maximum output settings during five cycles of the deflection voltage. High-speed spiral sweeps may be produced by replacing R_1 by a valve amplifier whose gain is swept through the required range by an auxiliary control circuit. Such a circuit is shown in Fig. 16-54 in which a pentode valve is used as the control circuit. A sinusoidal voltage of constant amplitude and the desired sweep frequency is applied to the control grid of the valve. The amplified sine wave of voltage developed at the anode is applied across R_4 and C_3 in series. This network supplies the out-of-phase deflection voltages to the c.r.t. The amplification of the valve is swept through a definite range of values by a sawtooth waveform of voltage applied to the screen grid. This causes the amplitude of the sweep voltage to increase and produces the spiral sweep as previously explained.

When a circular or spiral sweep is employed to measure an interval of time, some method must be used to indicate the beginning and end of the time interval. This may be done by applying a positive voltage pulse to the screen grid of the control valve. The basic circuit used to apply signal pulses to a circular sweep is shown in Fig. 16-55. Transmitter and echo pulses applied to the screen grid increase the gain of the amplifier momentarily. This increases the amplitude of the deflection voltages applied to both the X- and Y-plates. The trace jumps suddenly to a circle of larger diameter for the duration of the pulse as shown in the diagram.

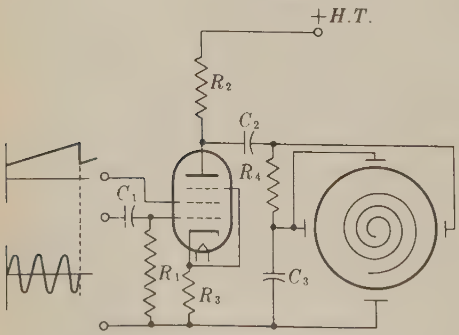


FIG. 16-54 Spiral-sweep generator circuit.

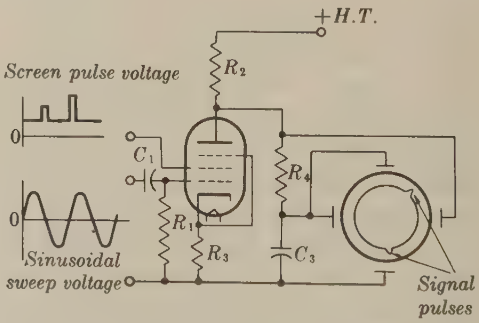


FIG. 16-55 Signal pulses on a circular sweep.

16.7 Questions and problems.

1. What are the three basic types of scans used in Service radar presentations?
2. Outline the method by which the range to a target is presented by a type *A* scan.
3. How is the bearing of a target determined when a type *A* scan is used?
4. Why is it desirable to have a linear sweep in the c.r.t. circuit?
5. What information is presented by a type *B* scan?
6. Explain how the aerial is synchronized with the horizontal sweep in a type *B* presentation.
7. (a) What is the meaning of the initials *PPI*?
   (b) What letter is used to designate this type of scan?
8. How does the bearing presentation in a *PPI* system differ from that in a type *B* system?
9. (a) What is meant by the term long persistence of vision?
   (b) Why is it desirable in the case of *PPI* presentations?
10. What is the principal merit of the *PPI* scan?
11. (a) What is a thyratron?
    (b) Why is it sometimes called a switching valve?
12. Explain the switching action of the thyratron in the sweep circuit of Fig. 16-5.
13. In what respect is the triode cut-off type of circuit preferable to the thyratron type as a radar sweep generator?
14. (a) Make a diagram of the triode cut-off sweep generator and draw the input and output voltage waveforms.
    (b) Explain the operation of the circuit.
15. (a) Make a diagram of a constant-current-pentode sweep generator.
    (b) Why would a triode valve not be suitable in place of the pentode?
16. How does the constant-current characteristic of the valve contribute to the linearity of the sweep?
17. (a) What controls the duration of the sweep in the above circuit?
    (b) What controls the speed of the sweep?
    (c) What controls the amplitude of the sweep voltage?
18. What is the purpose of a cathode follower feedback circuit following a sweep generator?
19. Explain how the cathode follower fulfils its purpose.
20. What factors limit the linearity obtainable with a cathode follower feedback circuit?
21. Make a diagram of a simple integrating circuit.
22. What is the effect of integrating
    (a) a square wave of voltage
    (b) a sawtooth wave of voltage?
23. How is integration employed to improve the linearity of a sweep generator?
24. Make a diagram of a Miller time-base generator and explain its operation with the aid of drawings of the voltage waveforms at the suppressor, grid and anode of the valve.

25. (a) What is a paraphase amplifier?
    (b) Why is such a circuit usually employed in radar indicators?

26. (a) Make a diagram of a two-stage resistance-capacitance coupled phase inverter.
    (b) How does it differ from a multivibrator circuit?

27. How does a split-load amplifier compare with a two-stage phase inverter as to
    (a) complexity of components
    (b) amplification?

28. What is the chief source of distortion in the phase inverter circuit?

29. How is this distortion largely overcome by the self-balancing phase inverter?

30. How does the second valve of the cathode-coupled phase inverter receive its signal?

31. How does the cathode coupling affect the gain of the first stage in the above circuit?

32. Explain how the cathode-coupled phase inverter may be used to supply d-c centering voltages as well as sweep voltages to the c.r.t.

33. State two methods used to deflect the electron beam horizontally and vertically by magnetic deflection in a c.r.t.

34. Which of the two magnetic deflection methods is commonly used with
    (a) *B* type scans
    (b) *PPI* scans?

35. What waveform of voltage must be used to produce a linearly increasing current in
    (a) a resistor
    (b) an inductance
    (c) an inductor containing some resistance?

36. Draw the basic circuit of a trapezoidal generator and explain its operation.

37. What is the purpose of the sweep power-amplifier stage?

38. Explain the operation of the power amplifier
    (a) between sweeps
    (b) during the sweep.

39. (a) What is the purpose of the diode connected across the deflection coil?
    (b) Explain its operation during the sweep and at the end of the sweep.

40. What determines
    (a) the duration of the sweep
    (b) the speed of the sweep
    (c) the physical length of the trace in the *PPI* tube?

41. Why must the amplitude of the jump voltage be changed for different ranges?

42. (a) On which range must the jump voltage be greatest?
    (b) How is this accomplished?

43. What is the purpose of the feedback introduced by the cathode circuits of V_2 and V_5 in Fig. 16-30?

44. Explain the operation of these two feedback lines.

45. What is meant by calibration of a scan?

46. Why is the period 12.2 microseconds commonly used as the basic unit of time in calibration circuits?

47. Explain with the aid of a diagram how a type A scan may be calibrated with 2000-yard and 10,000-yard markers.
48. Why must the calibration markers be synchronized with the beginning of the sweep? How is this accomplished?
49. What is the function of a gating circuit?
50. Make a diagram of a gated amplifier and explain its operation
    (a) as an open gate
    (b) as a closed gate.
51. Describe the appearance of the PPI calibration.
52. How is the PPI calibration pattern obtained?
53. What is the purpose of the A scan range step?
54. How is the position of the range step controlled
    (a) physically
    (b) electrically?
55. Describe the connection between the range step, handwheel and range dial.
56. What is the procedure for ranging on a target when a range step is used?
57. Make a diagram of a basic type of delay multivibrator and explain its operation.
58. Why is the basic type of multivibrator unsuitable as a range step generator?
59. Make a drawing of the phantastron circuit in which the pentagrid valve is represented as two separate valves to show its similarity to a multivibrator.
60. Which of the grids controls the division of current to the screen and anode of the valve?
61. Explain the voltage changes at the anode, grid 1, and cathode of the valve during the linear run-down period.
62. What causes the second switching action to take place in the phantastron?
63. What determines the duration of the pulse in the phantastron circuit?
64. How is the delay varied? Under what conditions is the delay a maximum?
65. What is the purpose of the cathode follower connected to the basic phantastron circuit?
66. What is the purpose of the diode connected in the time control circuit of the phantastron?
67. List four variations of the range step used in accurate ranging indicators.
68. How is a strobe spot produced? How is it applied to the c.r.t.? Describe its appearance.
69. What is the purpose of an expanded sweep in an indicator?
70. Make a diagram of an expanded-sweep generator.
71. Draw a c.r.t. scan to show the effect of the expanded sweep on echo pulses and calibration markers.
72. What is the purpose of a double sweep in a c.r.t.?
73. Make a block diagram of a circuit which produces a double sweep in an electrostatic deflection c.r.t.
74. Explain how a circular sweep may be produced in a c.r.t. employing electrostatic deflection.
75. In what two ways may the deflection voltages be altered to produce an ellipse?
76. How may a spiral sweep be produced?
77. Draw a diagram of a valve circuit used to produce a circular sweep and present signals on the trace.
78. Explain the operation of the circuit when a signal voltage pulse is applied to it.

CHAPTER XVII

RADAR SYSTEMS

Radar systems which are used by the Service vary greatly in complexity and performance. They range from the comparatively simple sets used as warning devices against surface targets to the highly refined equipments required to search for distant ships and aircraft and to train armament against these targets. The principles on which all of these sets operate are, however, basically the same. Consequently the functional components of all radar sets are the same. The degree of refinement of these components is determined by their particular applications and is found in the individual components, not in the addition of new basic units.

17.1 Fundamental components of a radar set.

A. Block Diagram

The fundamental components of all radar systems are shown in the block diagram of Fig. 17-1 and are as follows.

1. THE TIMER (TRIGGER UNIT OR SYNCHRONIZER).

This unit synchronizes the indicator and the transmitter circuits. At regular intervals it produces a signal pulse which causes the sweep to start in the indicator. At the same time, or at a precise time later, the timer produces a signal pulse which causes the transmitter to send out a pulse of radio-frequency energy. The timer repeats its sequence of synchronizing pulses at the pulse repetition frequency of the radar set.

2. THE MODULATOR.

The modulator receives the synchronizing pulse from the timer and develops a high-voltage, high-power pulse which turns on the transmitter for a short interval of time.

3. THE TRANSMITTER.

This unit is a very-high-frequency, high-power generator of radio-frequency energy. It produces a radio wave of constant frequency and amplitude for

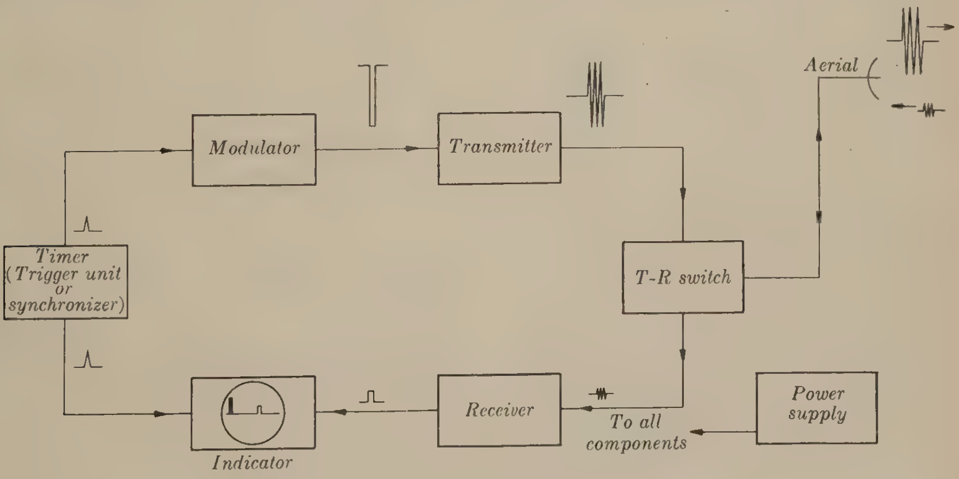


FIG.17-1 Block diagram of a radar set.

the short interval of time during which it is turned on by the modulator.

4. THE T-R SWITCH.

This is an electrically operated switch which effectively disconnects the aerial from the receiver and connects it to the transmitter during the production of the transmitter pulse. During the remainder of the operating period, the *T-R* switch connects the aerial to the receiver.

5. THE AERIAL.

The aerial system acts as a radiator of the transmitter energy when the transmitter is pulsed. It also serves to receive the echo pulses and convey them to the receiver. The aerial is usually made directive for both transmission and reception so that the bearing as well as the range of the target may be determined.

6. THE RECEIVER.

The receiver receives the r-f energy pulses reflected from the target. It amplifies these weak signals and produces video pulses of sufficient amplitude to produce visual indications in the indicator.

7. THE INDICATOR.

This unit is the electrical stop-watch which measures precisely the small interval of time required for the transmitter pulse to travel to the target and the echo to return. It may be used also to show the bearing and elevation of the target.

8. THE POWER SUPPLY.

The power supply delivers H.T. and L.T. voltages to all the valves, motor, etc. in the radar set.

B. Operation Factors

Before proceeding to a further discussion of the basic components of a radar set, it is well to consider certain other factors which affect the operation of the equipment. These include the choice of transmitter frequency, pulse repetition frequency and pulse width.

1. TRANSMITTER FREQUENCY.

The choice of carrier frequency is determined chiefly by the directivity required of the aerial system and the production, transmission and reception of the r-f energy. It is usually desirable to make the transmission as directive as possible in order to secure maximum range and accuracy of bearing. Moreover, the higher the frequency, the shorter the wavelength and so the smaller the physical size of the radiating elements which make up the aerial system. On the other hand, the difficulties involved in generating the carrier wave and in amplifying it in the receiver increase as the frequency increases. Special types of transmitting and receiving valves have been constructed with a minimum of interelectrode capacitance, stray inductance and transit time for use at radar frequencies. The range of radar frequencies now extends from about 100 Mc/s to 10,000 Mc/s.

2. PULSE REPETITION FREQUENCY.

This refers to the number of pulses per second sent out by the radar transmitter. It is determined largely by the maximum range of the radar set. Sufficient time must be allowed to elapse after a pulse has been transmitted to allow all echoes to return before a second pulse has been transmitted. Otherwise the reception and indication of echoes from distant targets may be confused by succeeding transmitted pulses. If, for example, a set has a maximum range of 200,000 yards, the time required for a pulse to travel to a target at maximum range and return is 6.1×200 or 1220 microseconds. In practice a period of 2000 microseconds would be allowed to elapse between transmitter pulses to avoid all possibility of confusion. The pulse repetition frequency of the set would then be 500 cycles per second.

3. PULSE WIDTH.

Pulse width refers to the duration of

the transmitter pulse. It is determined by the minimum target range. The transmitter pulse must end before an echo begins to return from the least distant target. Otherwise the returning echo is obscured by the transmitter pulse and the target echo is indistinguishable. The pulse width is usually of the order of one half to two microseconds. Sometimes the pulse width of a set may be changed by operating a switch on the front panel. The wider pulse is ordinarily used for distant objects as it contains more r-f energy, and the narrow pulse is used to follow a near target to minimum range.

4. Duty Cycle.

The pulse repetition frequency and pulse width determine another constant of the set called the *duty cycle*. The duty cycle is the ratio of the pulse width to the pulse repetition period. It may also be expressed in terms of the average and peak power. The average power is the power averaged over the pulse repetition period. The peak power of the radar transmitter is the r-f power radiated by the transmitter during the pulse. Since the transmitter is turned off for a relatively long time between pulses, the average power is a small fraction of the peak power. These two values of power are illustrated by the shaded portions of Fig. 17-2. The area enclosed by the pulses is equal to the area enclosed by the average power level and the zero line. Hence,

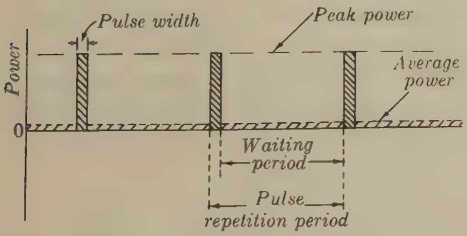


Fig.17-2 Peak and average power of a radar transmitter.

$$\frac{\text{average power}}{\text{peak power}} = \frac{\text{pulse width}}{\text{pulse repetition period}}$$
$$= \text{duty cycle.}$$

For example, if a radar set pulses for one microsecond every 2000 microseconds with a peak power of 100 KW, then the average power

$$= 100 \times \frac{1}{2000} = \frac{1}{20}\text{KW}$$
$$= 50 \text{ watts.}$$

$$\text{The duty cycle} = \frac{1}{2000} = .0005$$

$$\text{or, the duty cycle} = \frac{50}{100{,}000}$$
$$= .0005, \text{ as above.}$$

It is desirable to produce the highest possible peak power output in order to extend the effective range of the radar set to the maximum. The transmitter must, however, be operated within the limits of the average power dissipation. The duty cycle should therefore be made a minimum. This requires that the pulse width be kept small and that the repetition frequency also be limited to a small value.

In practice the pulse repetition frequency is kept up to about 500 pulses per second for motor-driven aerials to ensure that targets are not missed between pulses.

C. Long- and Short-Wave Radar

Although it is stated previously that the useful range of transmitter frequencies extends from 100 Mc/s to 10,000 Mc/s, Service equipments fall into two categories, long- and short-wave radar. Long-wave sets operate at about 200 Mc/s and are characterized by transmission-line oscillators with triode valves, coaxial-line feed to the aerial system and dipole aerials with linear directors and reflectors to improve directivity. Short-wave radar, also called *microwave radar*, is operated in the region between 3,000 and 10,000 Mc/s (10 cm. to 3 cm. wavelength). Microwave sets usually use

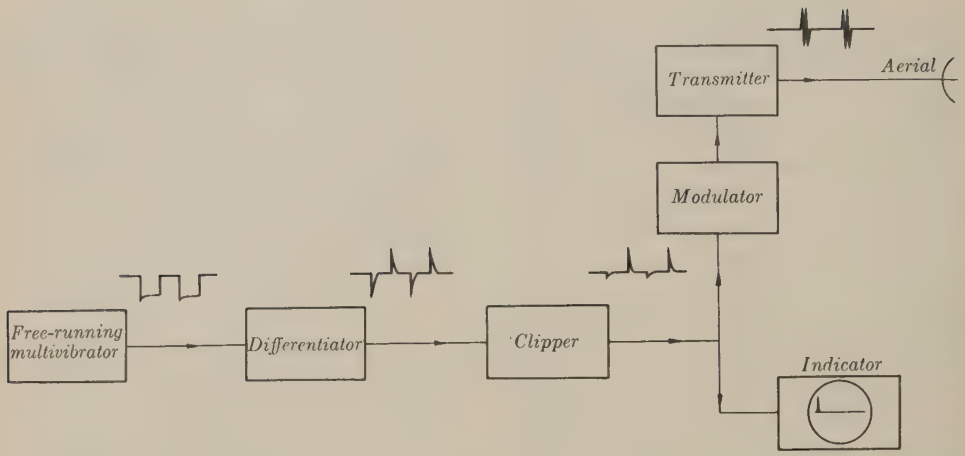


Fig.17-3 Block diagram of a separate timing unit.

magnetron transmitters, waveguide transmission systems and parabolic reflectors. The receiver local oscillator is generally of the klystron type and the first detector a crystal. Reference to these two types of radar sets will be made continually in the sections which follow.

17.2 Radar timers (trigger units or synchronizers). The timing unit which synchronizes the transmitter and indicator may be a separate component, as suggested in the block diagram, or it may be a part of the power supply, transmitter, or indicator. These four possibilities are discussed in the following paragraphs.

A. Separate Timing Unit

When a separate timing unit is used to control the radar set, a stable oscillator operating at the pulse repetition frequency is normally employed. It may be a multivibrator, blocking oscillator, or a sine-wave oscillator. Fig. 17-3 shows the block diagram of a typical multivibrator timing unit and the voltage waveforms produced. A free-running multivibrator produces a square wave of voltage at the required pulse repetition frequency of the set. This waveform is differentiated by an *RC* circuit to form a series of sharp positive and negative pulses. A diode clipper is then used to remove the negative pulses of the voltage waveform. The positive pulses are sent to the modulator and to the indicator simultaneously. Each of these pulses causes the modulator to turn on the transmitter and to start the sweep in the indicator at the same time. If a blocking oscillator is used instead of the multivibrator, the differentiator is not required. The sharp, single-swing oscillations are passed to the clipper stage and thence to the modulator and indicator units. If a sine-wave oscillator is used instead of a multivibrator, an overdriven amplifier is inserted ahead of the differentiator to develop a square wave of voltage. The use of a separate timing unit provides a certain amount of flexibility of the pulse repetition frequency. If several sets are operating simultaneously at the same place, they may be made to pulse at slightly different rates to prevent interference with each other.

B. Power-Supply Timing

In order to reduce the size and weight of transformers, capacitors and chokes in the power supply units, it is common

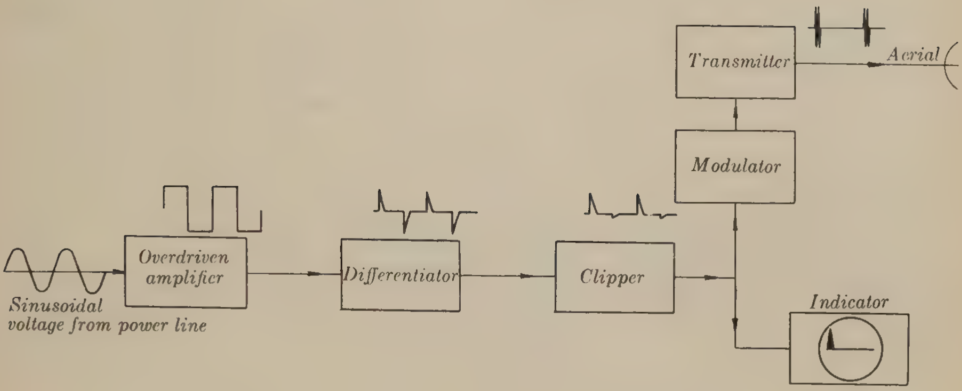


FIG. 17-4 Block diagram of a power-supply timer.

practice to use a power-supply frequency of the order of 500 cycles per second. This frequency is also a suitable pulse repetition frequency to control the radar set. A block diagram of the stages required to produce control signals from the line voltage is shown in Fig. 17-4. The sinusoidal voltage waveform from the supply line is applied to an overdriven amplifier which produces a square waveform of voltage. This waveform is then differentiated and clipped to form a series of positive pulses which trigger the modulator and indicator units. If several sets are to be operated at one time, the phase of the supply voltage is shifted by a different amount in the input to each set. For example, four sets may be operated one quarter of a cycle apart. If the supply frequency is 500 c/s, the period is $\frac{1}{500}$ of a second or 2000 microseconds. Four sets may therefore be pulsed at 500-microsecond intervals to provide 80,000-yard sweeps without interfering with each other.

C. Transmitter Timing

The transmitter may establish its own pulse repetition frequency and provide synchronizing signals to the other circuits in the set. A self-pulsing or blocking oscillator with properly selected values of circuit components may be used to provide the action. A part of the r-f transmitter energy is supplied to the indicator to start the sweep at the same instant as the transmitter is pulsed as shown in Fig. 17-5.

D. Indicator Timing

An alternative method of timing the radar set is by means of a crystal-controlled oscillator in the indicator unit. The calibration oscillator which operates at a frequency of 81.95 Kc/s to produce 2000-yard calibration markers is ordinarily used for this purpose. The frequency of the crystal oscillator is stepped down by a series of blocking oscillators or equivalent frequency dividers to obtain the desired pulse

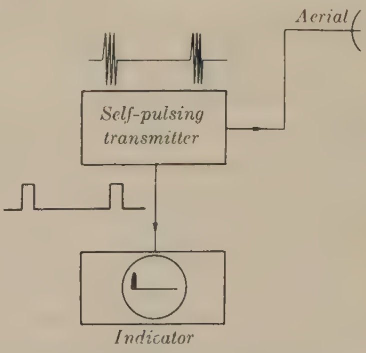


FIG. 17-5 Block diagram of a transmitter timing unit.

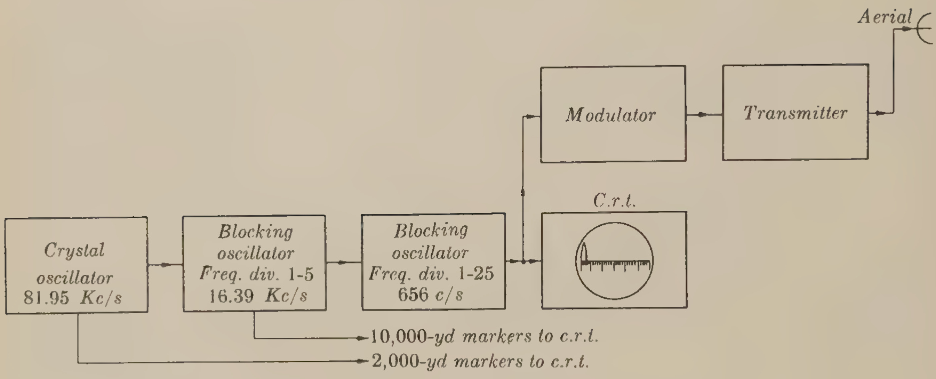

Aerial
Modulator
Transmitter
C.r.t.
Crystal oscillator 81.95 Kc/s
Blocking oscillator Freq. div. 1-5 16.39 Kc/s
Blocking oscillator Freq. div. 1-25 656 c/s
10,000-yd markers to c.r.t.
2,000-yd markers to c.r.t.

FIG.17-6 Block diagram of an indicator timing unit.

repetition frequency. The circuits required for this type of timing are shown in Fig. 17-6. The first blocking oscillator produces one cycle of output for every five cycles which it receives from the crystal oscillator. The output of the first blocking oscillator is supplied to the second blocking oscillator and to the c.r.t. as 10,000-yard markers. The second blocking oscillator produces one cycle of output for every 25 pulses which it receives from the preceding circuit. It therefore operates at a frequency of 656 c/s. These pulses are used to turn on the transmitter and to start the sweep in the indicator circuit. The pulse repetition frequency of the set is therefore 656 c/s.

17.3 Radar modulators and transmitters.

A. Transmission-Line Oscillator and Modulator

An example of a long-wave, low-power radar transmitter and modulator is shown in simplified schematic form in Fig. 17-7. The transmitter is a push-pull transmission-line oscillator with tuned circuits in the anode and cathode lines. The grids are connected to ground and the anodes supplied with 10,000 V H.T. through the charging resistor R_5 and a radio-frequency choke. The valves are maintained in a non-conducting state between pulses by a positive potential of 1000 volts applied to the cathodes from the anode of V_1. The grids are therefore 1000 volts negative to the cathodes and the storage capacitor C_3 is charged to 9000 V across the transmitter.

The modulator valve V_1 is also normally biased off by a positive potential applied to its cathode from the voltage divider R_2, R_4 across the H.T. supply. No current flows through R_1 and so the anode potential of the modulator is at the H.T. potential. The valve is turned on by a high-power positive pulse to the grid from a pulse-forming circuit which is not shown. This pulse is of rectangular waveform and of sufficient amplitude to drive the grid of V_1 considerably positive to its cathode. The valve conducts heavily and develops a large p.d. across R_1. The anode potential of V_1 therefore falls abruptly causing the cathode potential of the transmitting valves to fall also. This removes the paralyzing bias from the transmitter and oscillations are produced.

The duration of the transmitter pulse may be determined by either of two limiting factors. The storage capacitor C_3 supplies the cathode-to-anode current through V_2 and V_3. As this current continues to flow, C_3 discharges and the p.d. across it decreases. Eventually the p.d. which it applies to the transmitter

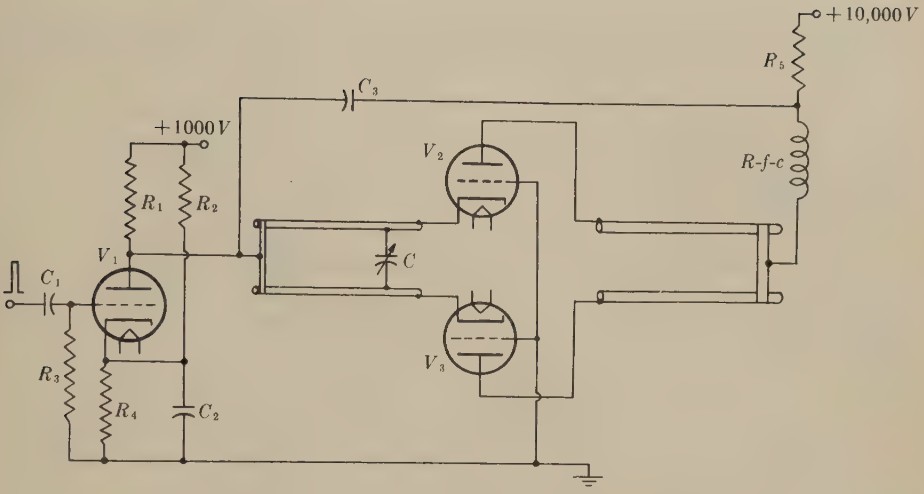


FIG. 17-7 Transmission-line oscillator and modulator.

falls below the critical value required to maintain oscillation, and the transmitter pulse ends abruptly. C_3 may be chosen of such a size as to permit the oscillation to continue for a period of one to two microseconds as required by the set. The capacitor then recharges gradually through R_5 before the circuit is triggered again.

If C_3 is made relatively large the duration of the transmitter pulse may be controlled by the modulator. When the trigger pulse applied to the grid of V_1 ends, the grid potential falls below cut-off again. The valve stops conducting and the anode potential again rises to 1000 volts. This re-imposes the paralyzing bias on the transmitter until the modulator is turned on again. If the modulator is used to control the duration of the transmitter pulse, the trigger unit which drives the modulator must produce a well shaped pulse. It must have steep leading and trailing edges and a flat top in order that the transmitter oscillations be of constant frequency and amplitude during the pulse.

The r-f energy developed by the transmitter is picked up by a tuned transmission-line section inductively coupled to the anode lines. The energy is fed by coaxial cable to the aerial assembly for radiation.

B. Magnetron Transmitter and Hard-Valve Modulator

A hard-valve circuit is shown in simplified form in Fig. 17-8. The circuit consists of a discharge-line pulse former, a cathode-loaded amplifier and a power-amplifier modulator. It is capable of supplying a high-voltage, high-power negative pulse to the cathode of the magnetron to turn it on each time a trigger pulse is received from the timer unit.

1. PULSE FORMER.

V_1 is a thyratron valve connected in a low-power discharge-line circuit which develops a rectangular voltage pulse. The valve is normally biased off by a negative voltage on the grid from a separate supply. The capacitors in the discharge line charge up through R_6 and R_4 to 300 volts from the H.T. supply. When a positive trigger pulse is applied to the grid of V_1 the valve conducts heavily. The impedance of the discharge line is matched to the impedance of R_3 and the valve when conducting. Consequently

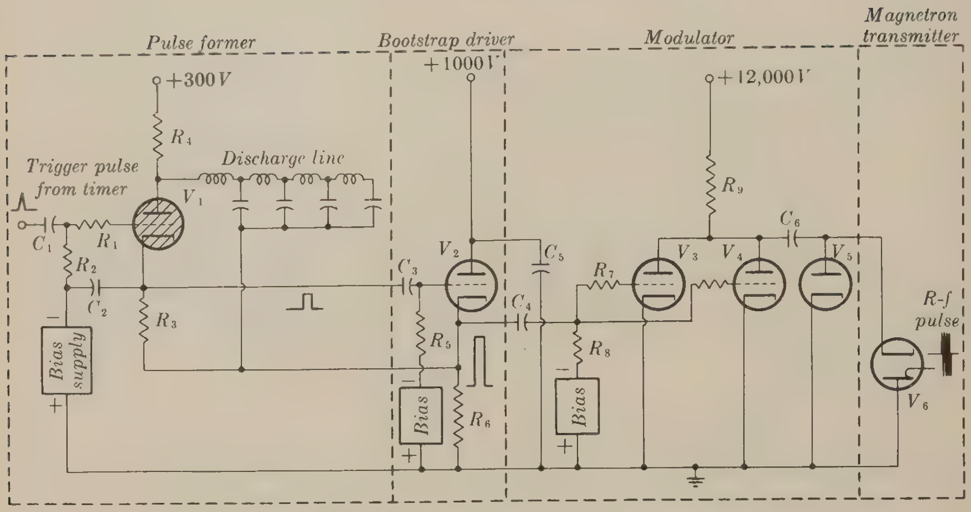


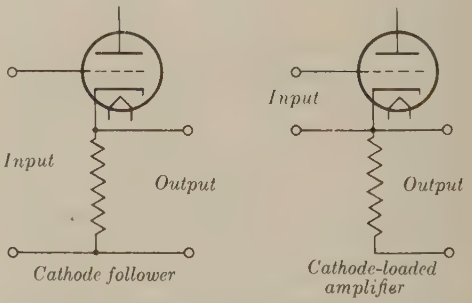
FIG. 17-8 Magnetron transmitter with hard-valve modulator.

the p.d. of the discharge line falls from 300 volts to 150 volts and a p.d. of 150 volts is developed across R_3 and V_1. This condition exists while a wavefront of current travels down the discharge line and returns. At the end of that time, the line is discharged and the p.d. across R_3 falls to zero. A rectangular positive pulse of voltage is therefore developed across R_3, with a duration of about one microsecond, the duration determined by the discharge line. This signal is applied between grid and cathode of the following valve.

2. DRIVER STAGE.

The driver is placed between the pulse former and the modulator to develop a high-voltage, high-power pulse of rectangular waveform to the modulator. It is necessary to drive the grids of the modulator valves considerably positive to their cathodes to make them conduct heavily enough to turn on the transmitter effectively. The driver shown in this circuit is a cathode-loaded amplifier, which is sometimes known as a *bootstrap circuit*. A cathode-loaded amplifier is one which has its load resistor in the cathode instead of the anode circuit. The input voltage is applied between the grid and cathode and the output is taken from the cathode load resistor. In this respect it is different from a cathode follower in which the signal is applied between the grid and the lower end of the cathode resistor (Fig. 17-9). The gain of a cathode follower can never be as great as one. The gain of a cathode-loaded amplifier is the same as that of an anode-loaded amplifier containing the same components.

Several operational features of this circuit which distinguish it from the more conventional amplifiers will become apparent when a typical cycle is discussed. V_2 is normally a much larger valve than V_1 and operates at a con-



Cathode follower

Cathode-loaded amplifier

FIG. 17-9 Cathode-follower and cathode-loaded amplifiers.

siderably higher supply voltage. The valve is normally biased off by a negative voltage on its grid. This is shown as a separate bias supply in Fig. 17-8. Capacitor C_5 therefore charges up to the H.T. supply voltage which is 1000 volts for this stage. No voltage exists across R_6 between pulses. At the time V_1 conducts, a positive voltage from R_3 is applied between grid and cathode of V_2. This overcomes the bias and causes the valve to conduct heavily. The 1000 V p.d. across C_5 is applied to V_2 and R_6 in series. Since the impedance of V_2 is small when conducting, most of the supply voltage is developed across R_6. Thus a positive voltage pulse of rectangular waveshape is applied from R_6 to the grids of the modulator valves V_3 and V_4. It lasts for the duration of the pulse produced by the discharge line and has an amplitude considerably greater because of the amplification of the stage.

A series of cathode-loaded amplifiers is well suited to the amplification of a positive pulse because all valves in the series receive positive input pulses and may be biased beyond cut-off during the long period between pulses. In this way the valves may be greatly over-driven during the pulse period without exceeding their average power-dissipation ratings. In pulse amplifiers of the more conventional anode-loaded type, each valve in the series reverses the polarity of the pulse. For this reason every second valve must be in a normally conducting state while the others are biased off between pulses. The maximum useful power output and over-all efficiency are limited by the power-handling capabilities of the normally conducting valves.

The bootstrap circuit derives its name from the fact that during the pulse period, the cathode of each stage rises in potential by an amount equal to the sum of all the cathode potentials which follow.

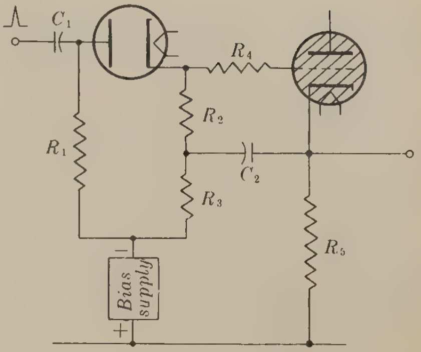


FIG. 17-10 Isolating diode in input circuit to a cathode-loaded amplifier.

For example, in the circuit of Fig. 17-8, the cathodes of V_1 and V_2 are at ground potential during the inactive period as no current flows through R_3 and R_6. If a p.d. of 750 volts is developed across R_6 during the pulse, and a p.d. across R_3 of 120 volts, then the cathode potential of V_1 must rise by 870 volts at the beginning of the pulse. The cathode potential of each stage in a series of cathode-loaded amplifiers rises in potential by an amount equal to the sum of all the voltages developed in the subsequent stages due to the bootstrap effect. This requires that the entire circuit surrounding V_1 be well insulated from ground. The input pulse to the grid of V_1 must be supplied through a circuit which is not directly connected to ground. One method of isolating the input circuit from the first valve of the bootstrap circuit is by means of a diode as shown in Fig. 17-10. The bias voltage to the grid of the thyratron stage is applied to both anode and cathode of the isolating diode. In the steady state, no current flows through R_1, R_2, R_3 and R_4 and so no p.d. exists across these resistors. The positive trigger pulse causes C_1 to charge through R_1 developing a positive voltage at the top of R_1. In parallel with this resistor is the diode valve in series with R_2 and R_3. Thus the anode of the valve is positive to its cathode. The valve conducts and the p.d. developed across R_2 is applied between grid and cathode of the thyratron. When the thyratron conducts and the bootstrap action takes place, the cathode potential of the thyratron rises sharply. This makes the cathode of the diode positive to its anode. The valve is in a non-conducting state and presents a high impedance which isolates the thyratron from the trigger circuit.

3. Modulator Stage.

The modulator stage (Fig. 17-8) consists of two large power amplifier valves V_3 and V_4 connected in parallel. V_5 is a charging diode and V_6 is the magnetron transmitter. The modulator is maintained in a non-conducting state by a bias arrangement shown here as a separate voltage source. When the supply voltage is connected, the capacitor C_6 charges through the anode load resistor R_9 and the diode. C_6 must therefore be capable of withstanding the full power-supply voltage of 12,000 volts. The cathode of the magnetron is only slightly positive to ground during the charging time due to the low impedance of the diode when conducting.

A positive pulse is applied to the grids of the modulator from the cathode circuit of the driver stage. It must be of sufficient amplitude not only to overcome the bias, but to drive the grids 100-200 volts positive to make the valves conduct heavily. The p.d. of the capacitor C_6 is applied across the magnetron and modulator in series. The p.d. required to make the modulator conduct is kept as low as possible as the power consumed by this valve is wasted as far as the magnetron is concerned. The capacitor C_6 discharges through the magnetron and the modulator valves. The capacitor must be capable of supplying the current required by the magnetron and maintaining the current at a reasonably constant value throughout the duration of the pulse.

Some conception of the operation of the capacitor may be obtained by consideration of an example in which the capacitance of C_6 is .025 microfarads and the pulse has one microsecond duration. Before the pulse, C_6 is charged to 12,000 volts by the power supply. All of this voltage appears between anode and cathode of the modulator valve which is biased off and is therefore an infinite impedance. The p.d. across the magnetron is zero. When the modulator grid is driven positive at the beginning of the pulse, the valve conducts with about 1000 volts between cathode and anode. The remainder of the p.d. across C_6 is applied to the magnetron, namely 11,000 volts. The initial current through the magnetron and modulator is about 11 amperes. As C_6 discharges throughout the pulse, the p.d. across it falls causing a reduction of current through the modulator and magnetron from 11 amperes to about 9 amperes. The average current flow is therefore 10 amperes, assuming a uniform decrease during the pulse. The amount of charge which must be supplied by C is given by the formula

$$Q = It$$

where Q = charge in coulombs
I = average current in amperes
t = duration of the pulse in seconds.

$$\text{Therefore } Q = 10 \times 1 \times 10^{-6} = 10 \times 10^{-6} \text{ coulombs.}$$

The fall in voltage, V, across C during the pulse is found from the formula

$$Q = CV$$

$$\text{Therefore } 10 \times 10^{-6} = .025 \times 10^{-6} V$$

$$\text{or} \quad V = 400 \text{ volts.}$$

Of this decrease in voltage, half appears across the magnetron and half across the modulator because their a-c impedances are approximately equal. Therefore as the voltage across the magnetron falls from 11 KV to 10.8 KV the current through it falls from 11 amperes to 9 amperes. A decrease in voltage across the modulator from 1000 volts to 800 volts corresponds to a decrease in current through that valve from 11 amperes to 9 amperes. Since the voltage and current changes are not great the power output and frequency of the magnetron is relatively constant during the pulse.

C. Magnetron Transmitter and Thyratron Modulator

An example of a thyratron modulator which may be used to supply a high-power negative pulse to the cathode of the magnetron transmitter is shown in Fig. 17-11. The modulator turns the transmitter on once during each cycle of the a-c supply voltage. Thus the pulse repetition rate of the set is the same as the frequency of the a-c supply. An explanation of each stage of the modulator-transmitter unit follows.

1. TRIGGER CIRCUIT.

V_1 and V_2, together with their associated circuit components make up a pulse-forming circuit which triggers the modulator. These two valves are supplied with out-of-phase voltages from the transformer T_2 secondary which is grounded at the centre. V_1 is a diode valve connected in a conventional half-wave rectifier circuit. When the upper end of the T_2 secondary becomes positive to ground, V_1 conducts and charges C_3 to the peak value of the secondary voltage. During this half cycle, V_2 does not conduct. The cathode of this valve is maintained at a steady positive d-c potential by the bias arrangement R_4C_2 and the grid of the valve is made negative by T_2 while V_1 is conducting.

During the opposite half of the supply cycle, the anode of V_1 is made negative by T_2 and this valve does not conduct. At the same time the grid of V_2 is made increasingly positive by the opposite end of the secondary winding. Near the peak of this voltage rise, the bias on the valve is overcome and the valve conducts heavily. C_3 discharges through R_5, C_2 and R_4 in parallel, and V_2. This develops a sharp rise in voltage at the top of R_5 which is the trigger pulse to V_4. At the same time the pulse of current through the valve re-establishes the bias across C_2. The valve quickly de-ionizes as the p.d. across C_3 drops and C_2 maintains the bias across R_4 until the next pulse is developed. The waveforms of voltage applied to the anode of V_1 and to the grid of V_2 by the power transformer T_2 are shown in Fig. 17-12 (a) and (b). C_1 and R_3 form a phase-shifting network which permits the voltage to the grid of V_2 to be shifted in phase from the voltage shown in the diagram. By adjustment of R_3 it is possible to operate several radar sets from the same a-c supply without their interfering with each other. Each set is pulsed at a different time during the supply cycle so that echoes from one radar set will not be received by another set while the sweep is taking place in that other set. The positive trigger pulses developed across R_5 which are applied to the grid of the modulator valve are shown in Fig. 17-12 (c).

2. MODULATOR.

V_3 and V_4 in the modulator operate in a manner similar to V_1 and V_2 in the trigger circuit, but with much higher values of voltage and current. T_1 is a high-voltage power transformer which supplies about 10,000 volts to the diode V_3 which is connected in a half-wave rectifier circuit. The anode of V_3 is made positive during the same half cycle of the power supply that V_1 is made to conduct. V_3 conducts to charge the capacitors in the discharge line to almost the peak voltage produced by T_1. During this half cycle the modulator valve V_4 is biased off and so presents an infinite impedance across the discharge line. Dur-

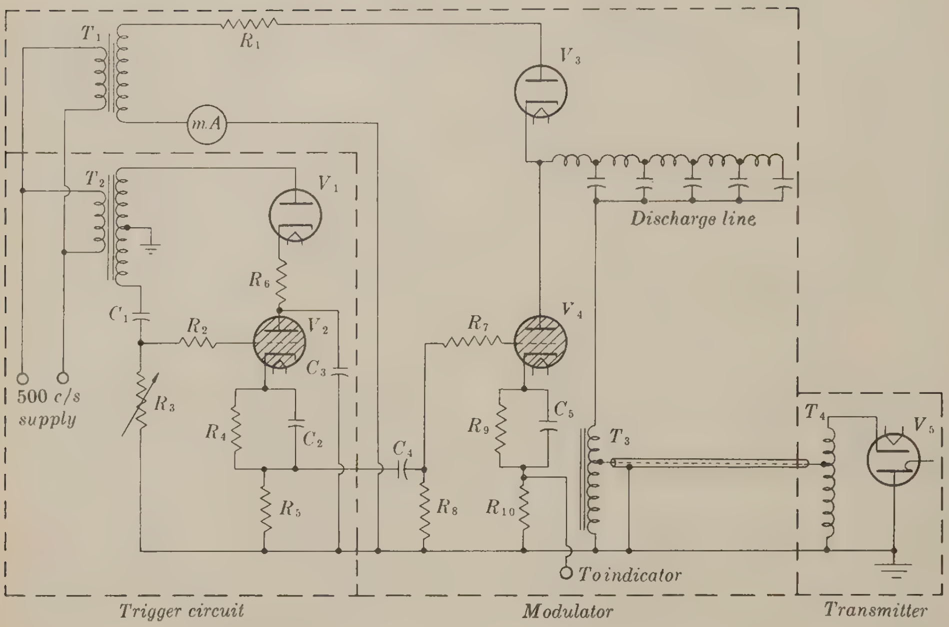


FIG. 17-11 Magnetron transmitter with thyratron modulator.

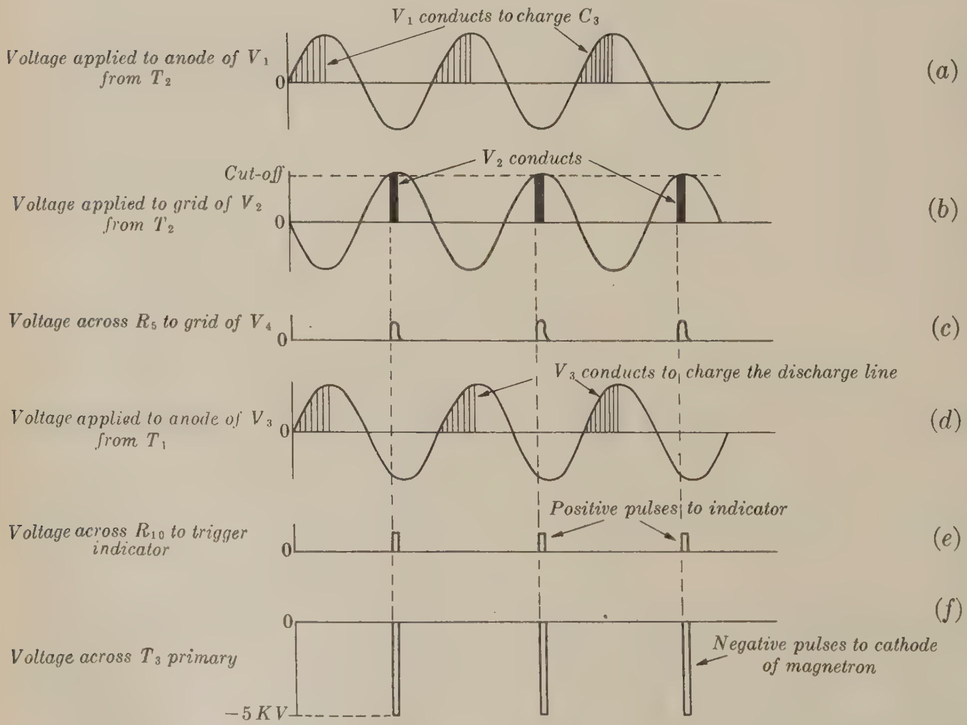


FIG. 17-12 Waveforms of operation of the circuit of Fig. 17-11.

ing the other half cycle of the supply voltage, the anode of V_3 is made negative and this valve does not conduct. The positive trigger pulse developed across R_5 in the trigger unit is applied to the grid of the thyratron V_4. This valve which is biased off for most of the cycle by the voltage across R_9 and C_5 is made to conduct heavily with a low impedance to the flow of current. The voltage of the discharge line is applied across the series network consisting of T_3, R_{10}, R_9 and C_5 in parallel, and V_4. The impedance of the discharge line is matched to the impedance of this network. The voltage to which the line is charged falls to half its previous value and the remaining voltage is impressed upon the network. Most of this voltage appears across the primary winding of the auto-transformer T_3 as the impedance of V_4 and its cathode circuit is relatively small. The duration of the pulse is controlled by the constants of the discharge line and is usually of the order of one half to two microseconds. At the end of this time, the capacitors in the line are discharged and the voltage across T_3 falls to zero. Since the electron flow is from the lower side of the line, down through T_3 and up through the cathode circuit and V_4, the potential developed at the upper end of T_3 is negative with respect to ground. The pulse also re-charges C_5 which maintains a bias voltage across R_9 between pulses. At the same time as the high-voltage pulse is produced across T_3, a positive voltage pulse is produced at the upper end of R_{10}. This pulse is used to trigger the sweep in the indicator at the instant the transmitter is pulsed. The waveforms of the voltages at the anode of V_3, at the top of R_{10} and across T_3 are shown in Fig. 17-12 (d), (e) and (f). The milliammeter connected in the ground return of T_1 measures the average direct current required to charge the capacitors in the discharge line. It is therefore also a measure of the average of the varying direct current which flows in the form of pulses through the primary circuit of T_3.

3. TRANSMITTER.

The voltage pulse produced across T_3 primary is stepped down by the auto-transformer action and applied to the inner conductor of the pulse cable to the magnetron. T_3 matches the impedance of the discharge-line circuit to the lower impedance of the pulse cable. At the transmitter end of the cable, a second auto-transformer T_4 steps up the pulse voltage to the cathode of the magnetron. The anode of the magnetron and the outer conductor of the pulse cable are at ground potential. Before the pulse takes place there is no p.d. across T_3 and T_4 and so the inner conductor of the pulse cable and the cathode of the magnetron are also at ground potential. During the pulse, the inner conductor of the pulse cable is made several kilovolts negative to the outer conductor. Due to the step-up action of T_4, the cathode of the magnetron is made about 12,000 volts negative to its anode and ground. This causes the transmitter to produce oscillations for the duration of the modulator pulse.

D. *Spark-Gap Modulator*

Thyratron modulators for magnetron transmitters have one serious drawback. The modulator valve must be a very large valve to withstand 12,000 volts between anode and cathode during the non-conducting period and to pass 10 amperes of current when conducting. The large filament current drawn by the valve requires that considerable heat be dissipated without the breaking of the glass seals. An alternative method of modulating the transmitter with a high-power discharge line is to use a spark gap instead of the thyratron valve. A simplified diagram of a spark-gap modulator is shown in Fig. 17-13, in which the simi-

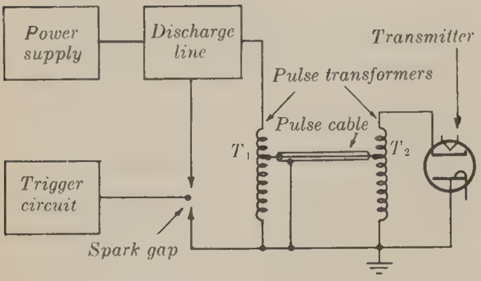


FIG. 17-13 Block diagram of a spark-gap modulator.

larity to the thyratron modulator is observed. The power supply charges the capacitors in the discharge line and provides the lower operating voltages to the trigger circuit. The trigger circuit supplies a pulse of voltage which causes the spark gap to conduct at a predetermined instant. The discharge line then discharges through the pulse transformer T_1 and the spark gap. The impedance of the discharge line is made equal to that of T_1 and the spark gap so that the line discharges in the form of a pulse. This pulse is carried by the pulse cable and T_2 to the magnetron causing it to oscillate and send out the transmitter signal. Two types of spark gap are in common use, the fixed and the rotary type.

1. FIXED SPARK-GAP MODULATOR.

The fixed spark gap shown in Fig. 17-14 consists of two spherical electrodes separated by a gap AB. This gap is made large enough so that it will stand the full H.T. voltage of the power supply without breaking down the air insulation. Mounted inside one of the spheres and insulated from it is a tungsten electrode C. The end of this electrode is separated from the sphere at A by a shorter distance than the main gap AB. A triggering potential is applied to C of sufficient amplitude to break down the air insulation between C and A. This produces a cloud of positive and negative ions which show as a spark at A. The positive ions are attracted by the high negative potential of the cathode, causing a break-down of the air between A and B. The ionization which is produced in this region forms a low-impedance path for the main discharge to take place from B to A. When the p.d. of the discharge line falls at the end of the pulse, the gas between B and A de-ionizes. The trigger pulse has also been removed and the discharge line begins a new charging cycle. The timing of the discharge can be precisely controlled by the application of the trigger pulse. The spark gap may be used for pulsing the magnetron with voltages several times as large as a thyratron modulator can stand. The current conducted by the gap during the pulse may be made to exceed 100 amperes.

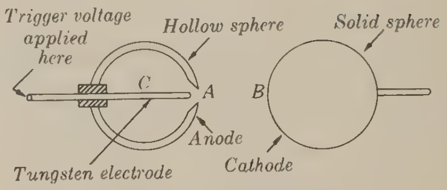


FIG. 17-14 Fixed spark gap.

2. ROTARY SPARK-GAP MODULATOR.

Nitrous oxide and ozone gases are produced by the high-tension discharge of the spark gap. These gases cause corrosion of the metallic surfaces of the spheres and tungsten electrode. This effect is more pronounced in the presence of water vapour and usually limits the life of the spark gap to about 200 hours. In order to extend the working period of the modulator, a rotary spark gap (Fig. 17-15) may be used in place of the fixed spark gap. A number of electrodes are mounted near the edge of a rotating metal wheel. They pass successively between a fixed electrode and a pointed ion generator which are mounted on the same bracket. The fixed electrode and point form one terminal of the spark gap while the rotating electrodes form the other terminal. The discharge between

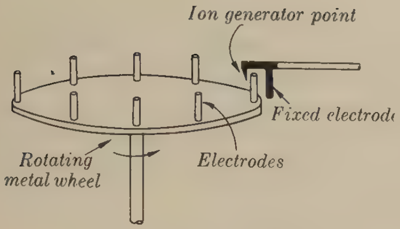


FIG. 17-15 Rotary spark gap.

the fixed and the rotating electrode takes place when the distance between them has decreased to a critical value. The rotating electrode passes somewhat closer to the point terminal than to the fixed electrode. This, together with the fact that the ion generator is pointed, causes the arc to begin between the rotating electrode and the point. Almost immediately the ionization and discharge spread to the gap between the fixed and the moving electrode causing the main discharge to take place. The charging of the discharge line is synchronized with the rotation of the wheel. The line is charged while the fixed electrode is opposite the space between two moving electrodes. The discharge then takes place each time one of the rotating electrodes passes between the ion generator and the fixed electrode. Since the corrosion caused by the arc is distributed over a number of electrodes, the life of the rotary spark gap is considerably longer than that of the fixed gap, being more than 1000 hours. It may be further extended by blowing the gases away with a fan or by absorbing them with a carbon absorber.

A spark gap itself acts as a high-frequency oscillator which radiates r-f energy. For this reason the gap must be enclosed in a grounded metal shield to prevent interference with other parts of the circuit. If this type of modulator is used in aircraft equipment, care must also be taken to maintain a constant air pressure around the gap as the breakdown voltage decreases when the pressure decreases.

17.4 Radar receivers. Improvement in the design and construction of radar receivers has contributed a great deal to extending the useful range of radar equipment, perhaps as much as any other single factor up to the present time. While the technician is seldom called upon to design such equipment an appreciation of its potentialities and of the critical nature of its adjustments is required if maximum efficiency of operation is to be maintained.

The purpose of the radar receiver is to accept small amounts of energy contained in the radar echo and to use them to control larger amounts of energy from the power supply. In this way the echo is amplified until it is of sufficient amplitude to cause a visual indication on the screen of the c.r.t. The echo pulse picked up by the receiving aerial consists of an electromagnetic wave of very short duration and very high frequency as determined by the radar transmitter. A particular example will serve to recall the nature of the receiver signal.

Suppose a microwave radar transmitter operates at a frequency of 3000 Mc/s and transmits a pulse of duration 2 microseconds. Then each transmitter pulse consists of a burst of r-f energy containing 6000 oscillations. This energy is radiated from the transmitting aerial in the general direction of a target. Of all the energy radiated by the transmitter, only a small fraction reaches the target, even with the use of a directional aerial. The energy that strikes the target is re-radiated in all directions. Consequently only a small percentage of the entire echo energy arrives back at the receiving aerial. It still contains, under ideal conditions, 6000 oscillations of constant frequency and amplitude but at a very low energy level. A high degree of amplification is therefore required of the

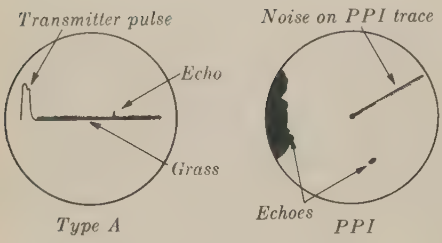


FIG.17-16 Appearance of noise on type *A* and *PPI* presentations.

receiver to amplify the pulse to the level required by the indicator.

The most important factor in the design and adjustment of the receiver for optimum performance is the signal-to-noise ratio.

A. Signal-To-Noise Ratio

While it is possible to secure any desired degree of amplification by connecting a sufficient number of amplifying stages in series, undesired voltages developed in the process may become so large as to mask the signal. These unwanted electrical disturbances constitute what is known as *receiver noise*. They appear on the type *A* c.r.t. as *grass* and on the *PPI* as an interrupted trace (Fig. 17-16). Noise generated in any stage of the receiver is amplified by all the stages which follow. Consequently, the noise produced in the initial stages is largely responsible for that appearing on the c.r.t. If the amplitude of the noise voltage generated in the first stage exceeds the amplitude of the echo signal, the signal voltage is lost. No amount of amplification by the succeeding circuits will serve to distinguish between signal and noise. It is desirable therefore to have the signal as large as possible and the noise level as low as possible. That is, the signal-to-noise ratio should be a maximum, and must be greater than one.

The various sources of noise produced in an amplifier are discussed in Chapter IX on communication reception. Those sources which are of prime importance in radar receivers are discussed again here, together with the methods that have been devised to reduce the effects to a minimum. They include thermal agitation, shot effect and induced voltages.

1. SHOT EFFECT.

Shot effect refers to irregular emission of electrons from the cathode of the valve. Even with steady d-c voltages applied to the valve electrodes, a varying current may flow from cathode to anode due to irregularities in cathode emission. A varying voltage is therefore developed across the anode load resistor which is amplified by all the succeeding stages. Shot effect is more pronounced in screen grid valves than in triodes. The presence of a positive screen grid to divide the electron stream between screen and anode circuits magnifies the shot effect due to random division. Consequently triodes are preferable to tetrodes and pentodes as amplifiers in the initial stages of the receiver. A valve with a high value of transconductance is also desirable as it exercises a greater control on the electron stream and therefore minimizes the effect of uneven cathode emission. An increased space charge about the cathode produced by increased heater potential is also effective in reducing shot effect in some valves.

2. THERMAL AGITATION.

Thermal agitation noise is caused by the random movement of the free electrons within a conductor. The average number of electrons entering and leaving a conductor over even a short period of time is the same. However, at any instant more electrons may be flowing in one direction in a conductor than in another direction, even when no p.d. is applied to the conductor. This results in a varying p.d. of random amplitude and frequency being developed by the conductor. It increases as the tempera-

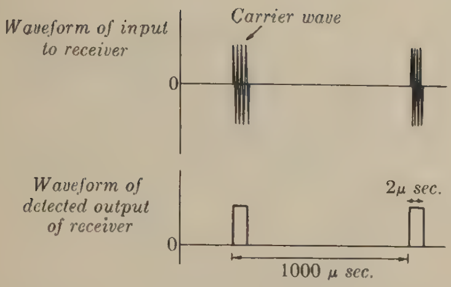

Waveform of input to receiver
Carrier wave
0
Waveform of detected output of receiver
2μ sec.
0
1000 μ sec.

FIG. 17-17 Waveforms of input and output voltages of a radar receiver.

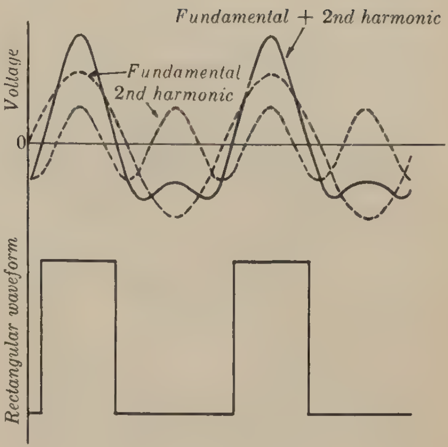


FIG. 17-18 Addition of fundamental and second harmonic in building up a square wave.

ture and resistance of the conductor increases. This effect is ordinarily of little importance in amplifiers of moderate gain but becomes a limiting factor in obtaining the highest possible signal-to-noise ratio in radar receivers.

3. INDUCED VOLTAGES.

Induced voltages are set up in the grid circuit of a valve by any irregularity in the electron stream flowing from cathode to anode. Shot effect and thermal agitation which are developed in a previous stage, induce grid voltages which further modulate the electron stream through the valve. The effect is magnified by the amplification of the stage. Since the induced effect is greater when the current through the valve fluctuates more rapidly, there is a greater amplification of the high-frequency noise voltages than the low-frequency noise voltages. Induced grid voltages therefore depend upon the mean frequency at which the receiver is designed to amplify. This becomes a factor which influences the choice of the intermediate frequency of the receiver.

B. Receiver Bandwidth

The noise voltages which are generated by shot effect, thermal agitation and induced grid voltages are distributed throughout the entire range of the frequency spectrum. One method of limiting their effect, therefore, is to limit the amplification of the receiver to a narrow band of frequencies which includes the transmitter frequency. This, however, is in direct conflict with the requirement for optimum shape of the rectangular waveform of the pulse outline. In order to prevent distortion of the pulse waveform, the receiver must be capable of amplification over a wide range of frequencies. This range includes the carrier frequency and the upper and lower sidebands. The sideband frequencies are equal to the carrier frequency plus and minus the frequencies contained in the modulation envelope. An examination of the rectangular outline of the detected echo pulse is required to determine these frequencies. The waveforms of the input pulses and of the modulation envelope are shown in Fig. 17-17. These waveforms cannot conveniently be drawn to scale. The pulse width is only $\frac{1}{500}$ of the pulse repetition period of 1000 cycles per sec. and the pulse should contain 6000 cycles for a carrier frequency of 3000 Mc/s. The question of analyzing the rectangular waveform into sinusoidal components of the proper amplitude and phase relationships has been referred to elsewhere in connection with symmetrical square and sawtooth voltage waveforms. The diagram of Fig. 17-18 shows the

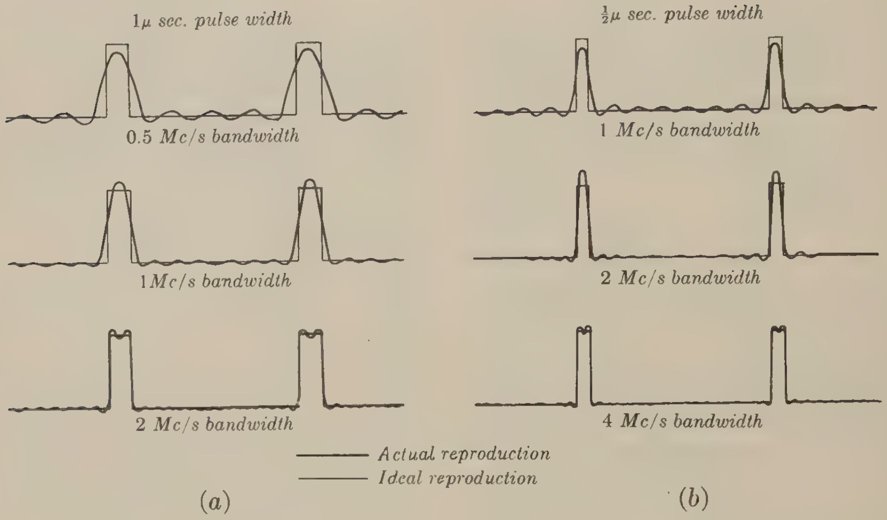


Fig.17-19 Receiver bandwidth requirements for narrow pulses (a) one microsecond duration (b) one half microsecond duration.

fundamental, the second harmonic and their sum which begins to resemble the rectangular waveform drawn below them. The addition of many more harmonics is required to improve the waveshape to a satisfactory degree. An amplifier must provide uniform amplification at all these harmonic frequencies if the rectangular waveform is not to be distorted. The quality of the reproduction increases as the amplifier bandwidth increases. Also, in order to make an amplifier yield a given quality of reproduction, the bandwidth must increase as the duration of the pulse decreases. This point is illustrated by the waveforms of Fig. 17-19 (a) and (b). The graphs of Fig. 17-19 (a) show the degree to which an amplifier will amplify a rectangular pulse of one microsecond duration when the bandwidths are respectively 0.5 Mc/s, 1 Mc/s and 2 Mc/s. Fig. 17-19 (b) shows a similar set of waveforms for a pulse of ½ microsecond, or half the width of those in Fig. 17-19 (a) when the bandwidths are respectively 1 Mc/s, 2 Mc/s and 4 Mc/s. In order to obtain the same quality of reproduction the bandwidth must be doubled, when the pulse width is halved. The diagrams of line 3 indicate the minimum requirements of receiver bandwidth for satisfactory reproduction. For a transmitter pulse duration of 1 microsecond the minimum permissible bandwidth is about 2 Mc/s. If the transmitter pulse is reduced to ½ microsecond, the receiver bandwidth must be extended to 4 Mc/s. For satisfactory reproduction of pulses of less than 1 microsecond duration the receiver bandwidth is usually in the neighbourhood of 8 Mc/s.

C. Receiver Block Diagram

A block diagram of a radar receiver is shown in Fig. 17-20. It contains the same stages as a communications receiver with the exception of the video amplifier in place of the audio amplifier. There are, however, extensive differences between the two receivers in the valves and circuits since radar receivers are required to amplify voltages at extremely high frequencies. The number of valve stages represented by each block of the diagram varies somewhat with the radar

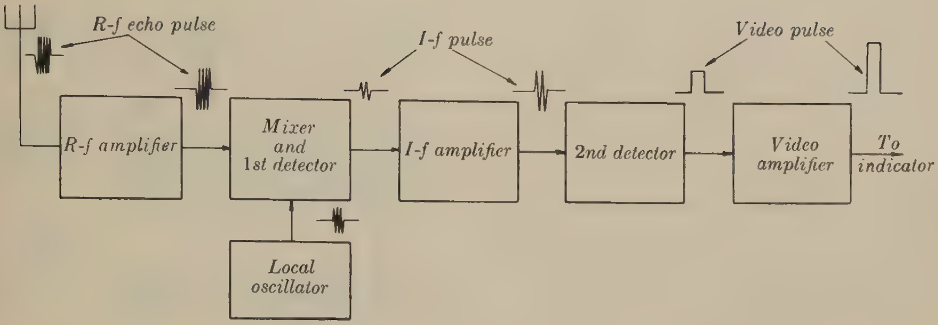


FIG.17-20 Block diagram of a radar receiver.

transmitter frequency and the intermediate frequency of the receiver. A set of examples of input frequencies, types of valves and numbers of stages for a long-wave and a microwave radar receiver are given in the table at the bottom of this page.

No attempt is made to amplify echo voltages at microwave frequencies. The echo pulses are fed directly to the mixer and first detector stage together with the local-oscillator signal. The output of this stage consists of pulses at the lower intermediate frequency. These are supplied to the intermediate-frequency amplifiers where most of the amplification takes place. A number of receiver stages used in long-wave and microwave radar are described in the following paragraphs.

D. R-F Amplifiers

The difficulties encountered in producing any useful amplification at frequencies above 100 Mc/s are discussed in Chapters II and IX. Noise considerations, reactance of valve leads, and instability due to interelectrode capacitance require that special valves and amplifier circuits be used to obtain any worthwhile gain of signal voltage without generating excessive noise voltage. The lighthouse triode and its associated circuit shown in Fig. 17-21 are suitable for amplifying voltages at frequencies up to 500 Mc/s. The valve has a standard base through which connections are made to the heater and cathode. Connections to the anode are made by way of the

STAGE	LONG-WAVE RADAR RECEIVER		MICROWAVE RADAR RECEIVER	
	INPUT FREQ.	NO. AND TYPE OF VALVE	INPUT FREQ.	NO. AND TYPE OF VALVE
R-F AMPLIFIER	200 Mc/s	2 Lighthouse triodes	3000 Mc/s	None
LOCAL OSCILLATOR	170 Mc/s	1 Acorn pentode	3060 Mc/s	Reflex klystron
1ST DETECTOR	200 Mc/s 170 Mc/s	1 Diode	3000 Mc/s 3060 Mc/s	Crystal
I-F AMPLIFIER	30 Mc/s	4 Pentodes	60 Mc/s	5 Pentodes
2ND DETECTOR	30 Mc/s	1 Diode	60 Mc/s	1 Diode
VIDEO AMPLIFIER	Rectangular Pulse	2 Pentodes	Rectangular Pulse	2 Pentodes

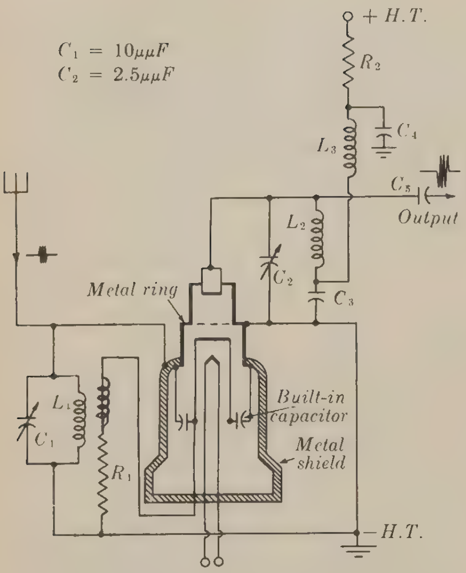


Fig. 17-21 R-f amplifier for frequencies up to 500 Mc/s.

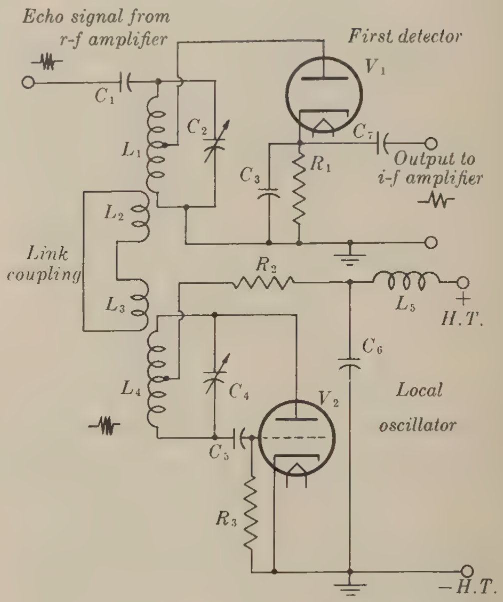


Fig. 17-22 Local-oscillator and first-detector circuits.

cap at the top of the valve and to the grid by means of a metal ring or disc about the tube midway between cathode and anode. The disc-seal construction and the reduced dimensions of the electrodes keep the interelectrode capacitance to a minimum.

The grid is connected to ground and the r-f signal from the aerial is applied to the cathode shell which is coupled to the cathode by a built-in capacitor. This causes the electron current from cathode to anode to vary at the frequency of the echo signal. The varying current through the valve sets up oscillations in the tuned anode circuit from which the output is taken. The variable capacitors C_1 and C_2 are adjusted to tune the grid and anode circuits to resonance at the echo frequency. The small values of capacitance employed in these circuits indicate the extremely high frequency of operation. Bias is developed across R_1 in the grid-cathode circuit. C_3, L_3 and C_4 constitute a decoupling network, C_5 is the coupling capacitor to the next stage and R_2 is an anode voltage dropping resistor. Since the grid is connected to ground this circuit is known as a *grounded-grid amplifier.*

E. Local Oscillator and First Detector

An example of a local oscillator and first detector used in a long-wave radar receiver is shown in Fig. 17-22. The echo signal from the r-f amplifier is coupled via C_1 to the resonant circuit L_1C_2 which is tuned to the frequency of the r-f carrier wave. Also coupled to this circuit is the output of the local oscillator. This circuit is a conventional series-feed Hartley oscillator which uses an acorn triode valve V_2. The local oscillator is shielded from the other receiver stages and coupling to the detector is provided by the link L_2L_3. Oscillations are therefore set up in L_1C_2 at two frequencies, the r-f carrier frequency which appears only when a transmitter or echo pulse is received in the aerial, and the continuous output from the

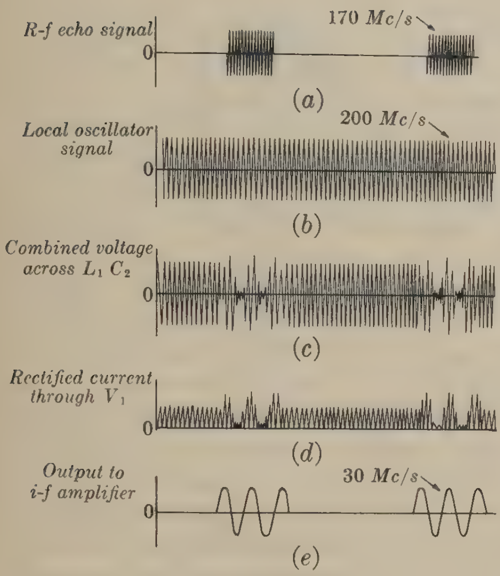


FIG.17-23 Voltage waveforms in first-detector stage.

local oscillator. The waveforms of these two signals are shown separately and in combination in Fig. 17-23 (a), (b) and (c) with given frequency values. The modulated wave is impressed upon the series diode circuit which rectifies the signal. The anode of the detector is tapped down on L_1 to reduce the loading effect of the diode on the tuned circuit. The rectified current flowing through R_1 develops the output voltage at the intermediate frequency which is the difference between the r-f and local oscillator frequencies. The waveforms of current through V_1 and of the output voltage are shown in Fig. 17-23 (d) and (e). The frequency of the local oscillator may be altered by means of the variable capacitor C_4. This is normally a front panel control. It is adjusted to keep the local-oscillator frequency far enough from the echo frequency to produce the required i-f. The i-f amplifiers produce their greatest gain when supplied with a signal at the frequency to which the circuits are pre-tuned. This condition is indicated by maximum amplitude of the echoes on the c.r.t. The setting of the local-oscillator tuning control is regularly checked during normal operation of the set to compensate for any changes in its frequency or in the frequency of the radar transmitter. This ensures that the receiver is kept operating at its maximum sensitivity.

The local-oscillator and first-detector circuits used in microwave radar receivers are considerably different from those employed in long-wave radar. The local oscillator is usually a reflex klystron of the McNally, Sutton or Shepherd-Pierce types as described in Chapter V. Valves of this type are constructed to operate as local oscillators with frequencies in the neighbourhood of 3060 Mc/s for 10 cm. sets and about 10,060 Mc/s for 3 cm. equipment. The frequency of the oscillator is adjusted by tuning plugs in the resonant cavity surrounding the valve electrodes. Fine control of the frequency for tuning the receiver is effected by a front panel control which sets the operating potential on the reflector electrode.

Valve detectors are not suitable for use at microwave frequencies because of their interelectrode capacitance and high noise level. For this reason crystals are universally used as first detectors. A schematic diagram of the aerial circuit, local oscillator, mixer unit and first detector is shown in Fig. 17-24. The mixer unit consists of a cylindrical cavity which contains a section of transmission line whose length is approximately one half of the transmitter wavelength. The upper end of the line is connected to the metal cylinder. The lower end of the line is by-passed to the cylinder for r-f current but is insulated from it for the intermediate frequency and d-c. The centre of the line therefore presents a high r-f impedance to ground. A crystal detector placed in a protective capsule is

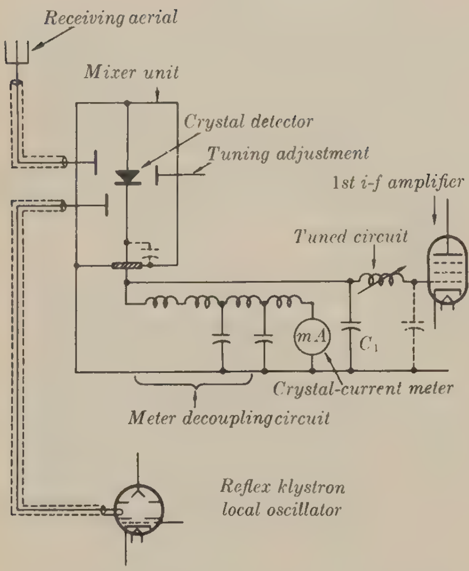
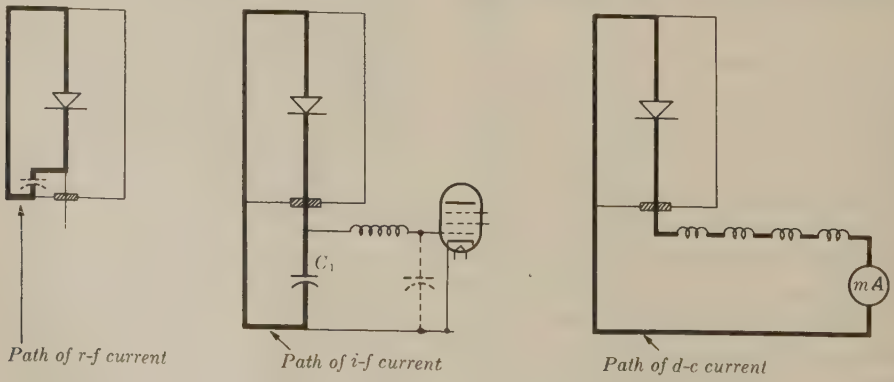


FIG. 17-24 Mixer unit and first detector for microwave radar receiver.

mounted near the centre of the line. The aerial signal is coupled to the line by capacitive coupling between the line and a disc at the end of the inner conductor of the concentric cable from the aerial. The degree of coupling is adjusted to prevent excessive loading of the crystal-aerial circuit. The local-oscillator signal is similarly coupled by a coaxial cable to a lower impedance point on the crystal line. The line is tuned by a pre-set variable capacitance between the line and the grounded cylinder. The crystal is therefore driven at two frequencies by the aerial circuit and the local oscillator. The crystal passes current in one direction and presents a high impedance in the other. Rectification of the mixed signal takes place and three components of current flow through the crystal. These components are r-f current at the carrier and local-oscillator frequencies, i-f current and direct current. They flow in separate paths as indicated by the thick lines in Fig. 17-25. The r-f current from the crystal is by-passed to the grounded cylinder through the capacitance of a mica washer between the line and the cylinder. The i-f current flows through the tuned grid circuit of the first i-f amplifier. This circuit is tuned to resonance at the intermediate frequency so that the i-f signal voltage is developed across C_1. The direct current flowing from the crystal passes through the inductors of the decoupling network, through the microammeter and the grounded cylinder, back to the other side of the crystal. The filter system prevents high-frequency currents from flowing through the meter. The crystal current is due mainly to the rectification of the a-c output of the local oscillator so that current flows through the meter even when no echoes are received. The meter is used in adjusting the output of the local oscillator to the level at which the



Path of r-f current

Path of i-f current

Path of d-c current

FIG. 17-25 Paths of r-f, i-f and d-c in first-detector circuit.

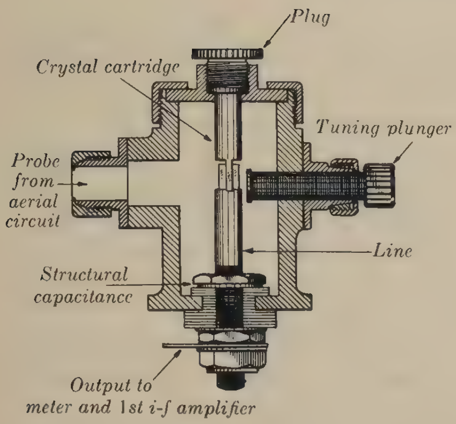

Plug
Crystal cartridge
Tuning plunger
Probe from aerial circuit
Line
Structural capacitance
Output to meter and 1st i-f amplifier

FIG.17-26 Cross-section of mixer and first-detector unit.

crystal operates most efficiently. It also serves as a check on the oscillation of the local oscillator during normal operation. A cross-sectional view of the structure of the mixer and detector unit is shown in Fig. 17-26. The local oscillator input does not appear as it is at right angles to the aerial input connector. An enlarged view of the crystal cartridge is shown in Fig. 17-27. The silicon crystal and cats-whisker are permanently mounted inside a metal cylinder for protection from shock and moisture.

F. Intermediate-Frequency Amplifier

The intermediate-frequency amplifier in the radar receiver usually consists of five or six stages similar to the one shown in Fig. 17-28. For long-wave radar, the entire amplifier strip may be installed in the main transmitter-receiver unit remote from the aerial. In microwave radar sets, it is desirable to have the transmitter, first detector and first two i-f amplifiers as close to the aerial as possible in order to keep the losses of r-f energy in waveguides and coaxial lines to a minimum. The initial two stages of the i-f strip then constitute the i-f preamplifier which is located on a separate chassis in the aerial assembly housing. The weak echo signals are therefore amplified before being fed by coaxial cable from the aerial assembly to the receiver-indicator chassis. The remaining stages of the i-f amplifier are located in the main receiver chassis together with the second-detector and cathode-follower circuits.

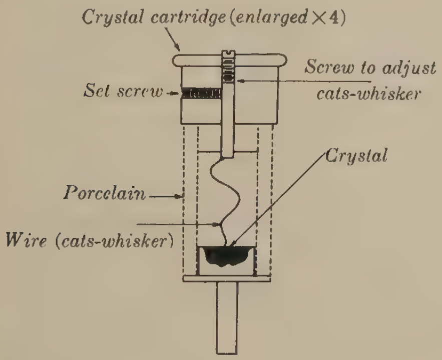


FIG.17-27 Structure of crystal cartridge for first detector.

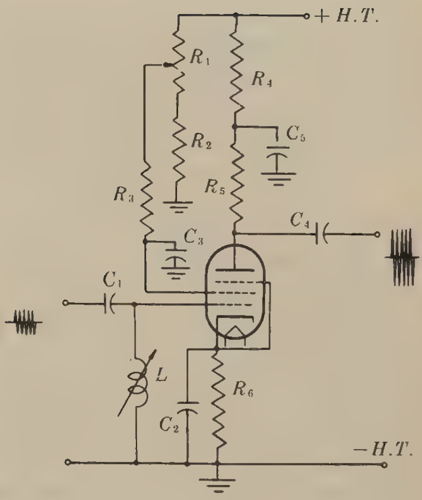


FIG.17-28 I-f amplifier for radar receiver.

The i-f amplifier is a tuned amplifier employing a high-transconductance pentode valve. The grid circuit is broadly tuned to the intermediate frequency, usually 30 Mc/s or 60 Mc/s. L resonates with the input capacitance of the valve to develop a large signal voltage between grid and cathode. An overall bandwidth of 4 to 8 Mc/s is obtained by broad tun-

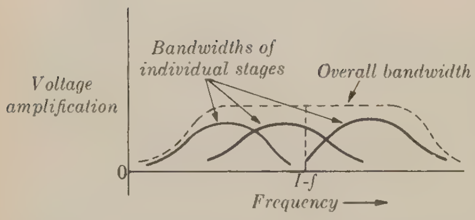


FIG.17-29 Staggered tuning in receiver stages to obtain wide band-pass.

ing of the individual grid circuits or by staggered tuning of successive stages. This method of obtaining a wide bandwidth with three stages of amplification is indicated by Fig. 17-29.

Resistance-capacitance coupling and small values of load resistance are used in the anode circuit (Fig. 17-28) to maintain the wide frequency response. Transformer coupling may be used between stages instead of *R-C* coupling. R_4 and C_5 constitute a decoupling network to prevent oscillation of the circuits due to feedback through the H.T. supply. Bias is provided by R_6 and C_2 in the cathode circuit. The gain of the amplifier is controlled by varying the potential on the screen grid by means of R_1. This resistor is part of a voltage divider R_1,R_2 across the H.T. supply. Alternatively, the gain may be controlled by varying the negative bias voltage to the control grid as variable-mu pentodes are generally used. In either case, the gain control is usually made to operate on the first two valves of the main i-f strip. This prevents overloading of the final stages and provides sufficient range of control over the height of grass and echoes on the c.r.t.

G. Second-Detector and Video Circuits

The amplified echo signal at the intermediate frequency is applied across the tuned inductor L to the anode of the second detector V_1 (Fig. 17-30). The valve conducts only on positive half cycles and passes three components of current. The intermediate frequency component is bypassed through C_2 to ground. The d-c and video components of current flow through R_1 producing a positive rectangular pulse of voltage each time an echo is received. This pulse is the modulation envelope of the echo signals. The waveforms of input voltage, current through the valve and output voltage are shown in Fig. 17-31.

The output voltage from the second detector is applied via the series resistor R_2 to the grid of the video amplifier V_2. The video amplifier is a resistance-capacitance-coupled circuit employing a

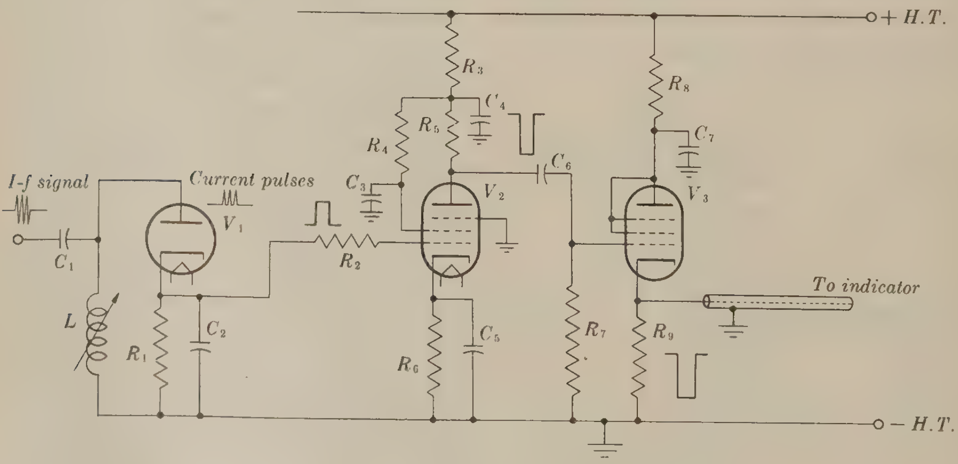


FIG.17-30 Second-detector and video circuits.

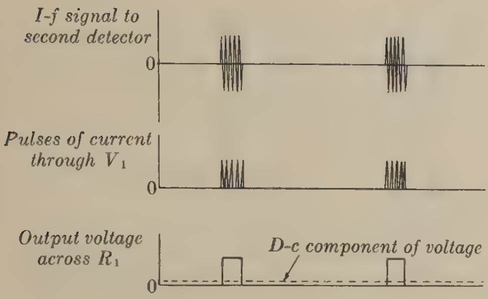


FIG. 17-31 Voltage and current waveforms for circuit of Fig. 17-30.

high-transconductance pentode valve and a small anode load resistance. In addition, series and shunt compensation networks are sometimes included in the grid circuit in order to improve the video-frequency response. Bias is developed in the cathode circuit by C_5R_6. The anode circuit is decoupled from the H.T. supply by R_3C_4. The positive input pulse causes a negative-going pulse to be produced at the anode of V_2. This pulse is applied to the grid of V_3.

The second video stage is a cathode-follower circuit for matching the high-impedance output of V_2 to the low impedance of the coaxial line. This line carries the signal from the receiver to the indicator chassis. The valve is normally a triode or a pentode connected as a triode. In the absence of a signal, the direct current flowing through the valve maintains the cathode considerably positive to ground. When a negative pulse is applied to the grid, the current through the valve decreases and the p.d. across R_9 decreases. Consequently a negative pulse is also produced at the cathode. The amplitude of the output voltage is always slightly less than the amplitude of the input applied to the grid. Although the gain of the stage is less than one, the low output impedance provides a means of feeding the signal through the low-impedance line with a minimum of loss. The negative-going pulses may be applied to the lower Y-plate of the c.r.t. in the indicator chassis to cause upward deflection of the trace. More generally, they are inverted by another video amplifier in the indicator chassis and applied as positive-going pulses to the upper Y-plate.

H. Receiver Blocking and Recovery Time

Care must be taken in operating the radar set to ensure that the receiver is not overloaded by excessively large signals applied via the aerial circuit. This most often occurs when the strong transmitter pulse is picked up directly by the receiving aerial or by reflection from nearby objects. A special protective device called a *transmit-receive* (*T-R*) *switch* is usually included between the receiving aerial and the receiver to protect the crystal first detector. Additional precaution is sometimes required in the video stages to prevent overloading. It may occur when a large signal drives the grid of the *R-C* coupled amplifier excessively positive. Grid current flows to charge the coupling capacitor. When the input pulse is removed, a negative voltage is applied to the grid until the coupling capacitor discharges through the grid-leak resistor. During this period the receiver may be biased off and become insensitive to echo pulses which follow. This effect is illustrated by the circuit of Fig. 17-32 and waveforms of Fig. 17-33. The circuit is a video amplifier at the end of the receiver chain. It is normally biased near cut-off by the positive potential on the cathode from the voltage divider R_4, R_5 and R_2 across the H.T. supply. The echo pulse causes a rise in grid potential which causes the valve to conduct and develops a negative pulse at the anode.

The type of signal under discussion is an excessively large pulse such as might come from the transmitter directly, followed by a pulse of normal amplitude.

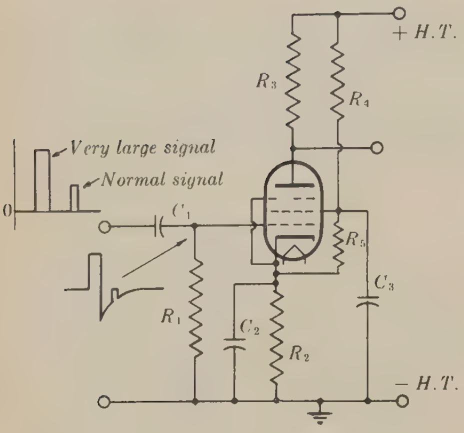


FIG. 17-32 Receiver video amplifier.

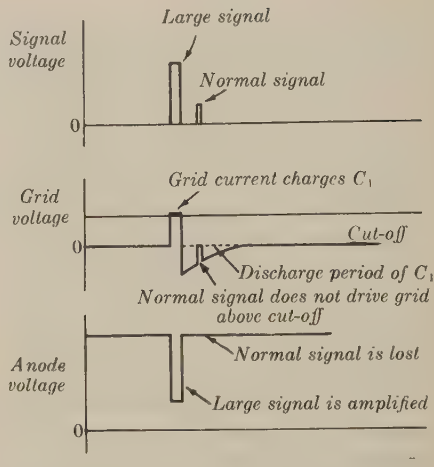

Large signal
Signal voltage
Normal signal
0
Grid current charges C_1
Grid voltage
Cut-off
0
Discharge period of C_1
Normal signal does not drive grid above cut-off
Anode voltage
Normal signal is lost
Large signal is amplified
0

FIG. 17-33 Receiver blocking due to large time constant in the grid circuit of video amplifier.

The large pulse not only causes anode current to flow through the valve, but also drives the grid positive to the cathode. As a result, grid current flows to charge C_1 by the amount that the signal voltage exceeds the cathode potential. At the end of the pulse, the grid voltage falls suddenly, but C_1 cannot discharge instantaneously. The grid potential drops below cut-off causing the anode potential to rise to H.T. The grid voltage rises exponentially toward cut-off as C_1 discharges through R_1. The duration of this period is known as the *recovery time* of the circuit and is determined by the time constant of R_1C_1. It is the time taken for the set to recover from the paralyzing effect of the large signal. During this period, the normal signal applied to the grid may be insufficient in amplitude to raise the grid potential to cut-off. The anode potential therefore remains constant and the normal signal is lost. The effect is illustrated by the waveforms of Fig. 17-33. This may limit the minimum effective range of the radar set as the receiver is silenced for a period following each transmitter pulse. Blocking may also occur as a result of a rise in cathode potential caused by a very large signal. In either case, the effect may be minimized by making the time constants of the grid and cathode circuits small. This cuts down the recovery time of the circuit and therefore reduces the possibility of echoes being lost in the receiver and not appearing on the c.r.t. trace.

More elaborate means of preventing blocking may be necessary in short-range surface and aircraft warning radar. The recovery time of the grid circuit may be reduced by connecting a diode in parallel with R_1 as in Fig. 17-34. The anode is grounded and the cathode connected to the upper end of R_1. The diode presents a high impedance across R_1 during the pulse. At the end of the pulse, the valve conducts with low impedance to discharge C_1 quickly to its

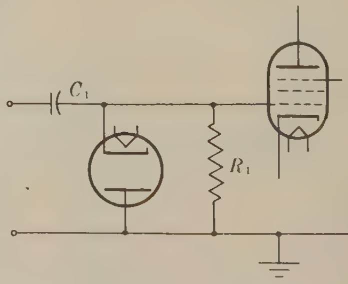


FIG. 17-34 Diode used to reduce recovery time in a video amplifier circuit.

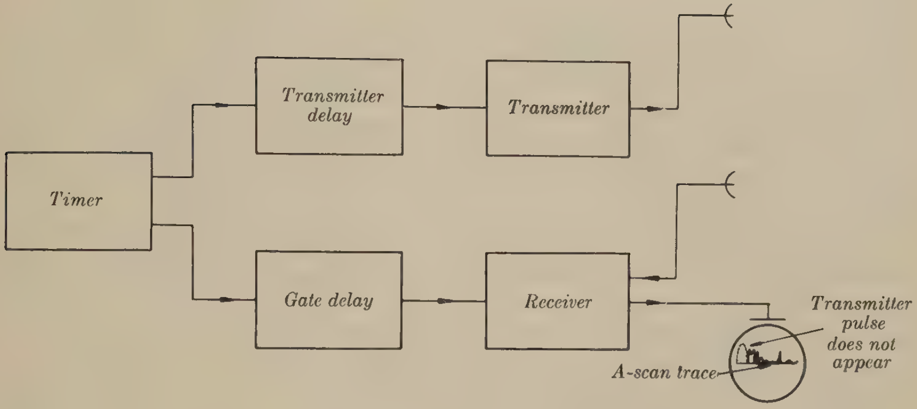


FIG. 17-35 Block diagram of receiver gating circuit.

normal voltage. Alternatively the signal voltage applied to the video amplifier may be limited to a value which will prevent blocking. This may be done by the preceding video amplifier which amplifies negative-going echo pulses. The bias on the valve is adjusted so that an excessively large signal drives the grid quickly beyond cut-off. The rise in anode potential is limited to a value which is not sufficient to drive the grid of the next valve positive to its cathode. In this way the flow of grid current is avoided and consequently blocking does not take place.

I. Receiver Gating

A more effective method of ensuring that the receiver is not paralyzed by the transmitter pulse is by gating the receiver off while the transmitter is turned on. Mechanical means of suppressing the receiver for the brief interval required are impractical. Consequently a gating circuit is used to bias off the receiver for a few microseconds beginning slightly ahead of, and ending just after the transmitter pulse. The exact circuits used to produce the gate voltage may vary considerably in detail, but the basic stages are suggested by the block diagram of Fig. 17-35 and the waveforms of Fig. 17-36. The delay circuits are usually multivibrators of the biased-off type. At time t_0 the timing unit sends a synchronizing pulse to the transmitter delay circuit and to the receiver gate delay circuit. At a predetermined time later, t_1, the transmitter delay cycle ends and this circuit turns on the transmitter. The receiver gate delay is adjusted to run until time t_2, the end of the transmitter pulse. The negative square wave of voltage produced by the gate delay is applied to the suppressor grid of one of the early r-f amplifier stages. It biases off this valve and renders the receiver inoperative for a period which includes

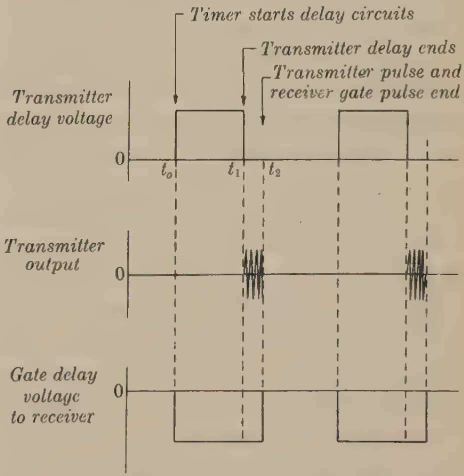


FIG. 17-36 Timing of gate voltage to silence receiver during transmitter pulse.

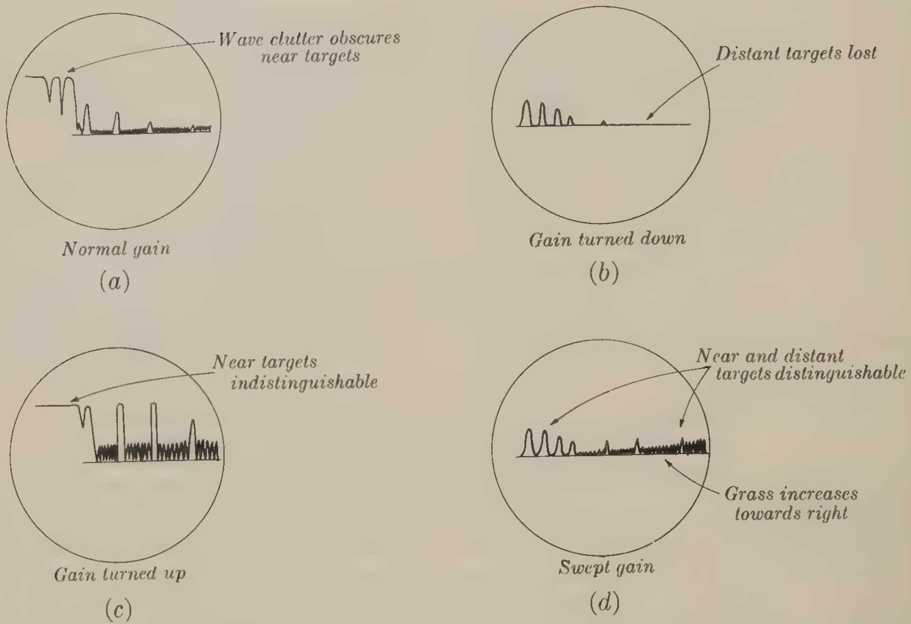

Wave clutter obscures near targets
Normal gain
(a)
Distant targets lost
Gain turned down
(b)
Near targets indistinguishable
Gain turned up
(c)
Near and distant targets distinguishable
Grass increases towards right
Swept gain
(d)

Fig.17-37 Wave clutter on type *A* presentation with (a) normal gain (b) gain turned down (c) gain turned up (d) swept gain.

the transmitter pulse. At the end of this time, the suppressor voltage is returned to normal and the receiver is free to receive echoes with no possibility of blocking by the transmitter pulse.

J. Wave Clutter

The signals received from nearby objects are very much stronger than those received from more distant objects. Moreover, in the case of shipborne radar, objects near the radar set, such as the ship's superstructure, ocean waves, rainstorms, etc., reflect a great deal of energy to the receiver. As a result, a large amount of deflection on the trace is usually seen at the beginning of the sweep. It is referred to as *wave clutter* and causes the c.r.t. pattern to appear as in Fig. 17-37 (a). If the gain of the receiver is turned down to make targets at close range distinguishable from the wave clutter as shown in Fig. 17-37 (b), the echoes near the right hand end of the trace do not appear. If the gain is turned up as shown in Fig. 17-37 (c) to make distant objects appear on the trace, the wave clutter at the left-hand side of the c.r.t. not only obscures echoes at close range, but may paralyze the receiver by over-loading. An ideal situation would be to turn the receiver gain down at the beginning of each sweep and to turn it up progressively as the sweep advances. The resulting pattern is shown in Fig. 17-37 (d). A circuit which performs this operation is called an *anti-wave-clutter* unit. It produces a varying voltage which alters the receiver gain from a minimum to a maximum as the sweep takes place. For this reason it is also called a *swept-gain* circuit.

A block diagram of a radar receiver employing an anti-wave-clutter unit is shown in Fig. 17-38. The unit receives two input voltages, the i-f signal from the pre-amplifier located in the aerial assembly and a negative pulse from the timer or modulator at the same time as the transmitter is pulsed. The anti-wave-

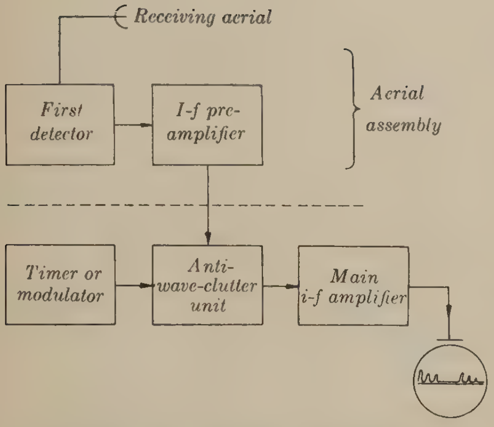


FIG.17-38 Block diagram of a radar receiver with anti-wave-clutter unit.

clutter unit develops a negative voltage waveform which reaches a maximum at the beginning of the trace and decreases exponentially toward zero as the sweep progresses. Two controls on the front panel of the unit make it possible to adjust both the duration of the waveform and the rate at which it approaches zero. This voltage is combined with the i-f signal voltage as indicated in the block diagram. The combined signal is applied to the grid of the first main i-f amplifier, a variable-mu valve. The valve is effectively biased down to a low-transconductance region of operation at the beginning of the sweep to reduce the gain of the stage. As the sweep progresses, the operating point of the valve shifts to a region of higher transconductance and the gain increases.

An anti-wave-clutter or swept-gain circuit is shown in Fig. 17-39. Before the sweep begins, there is no charge on C_1, no p.d. across any of the resistors and no current through the two diode valves. The input to the circuit is a negative square trigger pulse of voltage from the timer or modulator at the time the transmitter is pulsed. This voltage is applied across R_1 and the series circuit consisting of V_1, R_2 and C_1. The cathode of V_1 is made negative with respect to its anode, causing the valve to conduct. C_1 commences to charge through R_2 at a rate determined by the time constant of R_2C_1. The p.d. to which C_1 charges during the brief period of the trigger pulse is determined by the value of R_2 which is adjustable. This adjustment therefore controls the maximum amplitude of the negative swept-gain voltage. Its effect is illustrated by the waveforms of Fig. 17-40. If R_2 is made large, C_1 charges more slowly and the maximum voltage attained by C_1 during the charging period is less.

When the trigger pulse is removed, C_1 is prevented from discharging through V_1 as the diode will not conduct in that direction. The p.d. across C_1 is the input voltage to R_3 which is normally a much larger resistance than R_2. In parallel

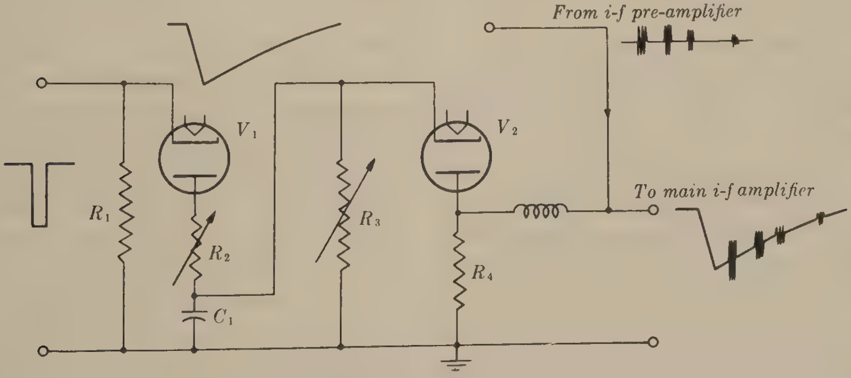


FIG.17-39 Anti-wave-clutter or swept-gain circuit.

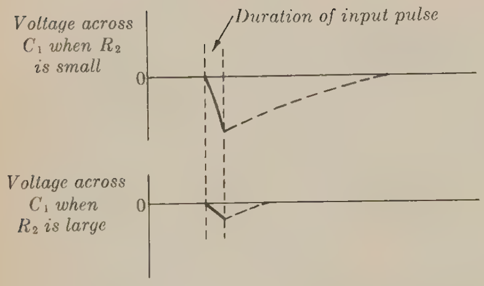


FIG. 17-40 Effect of varying amplitude control in the anti-wave-clutter circuit.

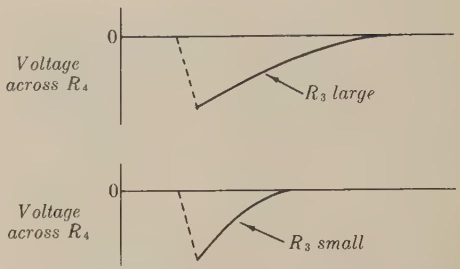


FIG. 17-41 Effect of varying duration control of anti-wave-clutter circuit.

with R_3 is the series circuit consisting of V_2 and R_4. The input voltage makes the cathode of V_2 negative with respect to its anode. This valve therefore conducts and develops a p.d. across R_4, negative at the upper end of R_4. As C_1 discharges through R_3 and V_2, the negative voltage across R_4 decreases. The rate at which C_1 discharges is controlled by the parallel resistor R_3. If this resistor is made small, C_1 discharges more rapidly through the two parallel paths and the second portion of the swept-gain voltage approaches zero more rapidly. The effect of varying R_3 with R_2 constant is indicated by the waveforms of Fig. 17-41. In practice, R_2 and R_3 are adjusted so that the echo signals are less than saturation level but greater than the grass level. The transmitter pulse does not show on the c.r.t. as the receiver is gated off during the time the transmitter is turned on.

Swept gain may be used with *PPI* presentations as well as with type *A* scans. The setting of the two controls to give best results with a *PPI* is usually different from that required by the *A* scan. If both presentations are in use at one time, separate swept-gain circuits for the two tubes are desirable. The painting of the *PPI* is more intense nearer the centre, even with equal echoes. By proper adjustment of the controls a reasonably uniform painting of targets at all ranges may be obtained.

K. Logarithmic Receivers

The anti-wave-clutter circuit described in the preceding section provides a varying voltage which varies the gain of the receiver as the sweep progresses. The receiver gain is made a minimum for echoes from near targets and vice versa. Its purpose is to prevent strong signals from near objects from saturating the receiver and to permit weak signals from distant objects to produce normal indication on the c.r.t. scan. An alternative method of achieving this result is to make the receiver most sensitive to weak signals and least sensitive to strong signals. This method avoids discrimination against small echoes at short range which do not appear on the scan when swept gain is used. It can be shown mathematically that the best results are obtained by using a receiver with a logarithmic response. That is, the output voltage of the receiver is proportional to the logarithm of the input voltage. This effect is achieved by the use of non-linear amplifiers in the receiver strip. The gain of the individual amplifiers is normal for small signals, but is reduced for large signals by reducing the effective impedance of the load circuit. A logarithmic amplifier for use in the intermediate-frequency section ot a radar receiver is shown in Fig. 17-42. The i-f signal is applied via the tuned input inductance L_1, to the grid of the pentode amplifier. The amplified output

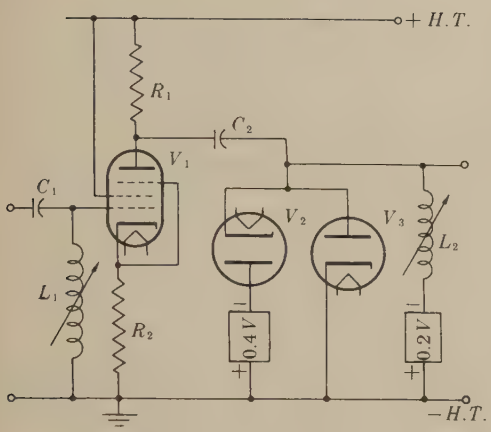


FIG.17-42 Logarithmic amplifier circuit for i-f stages.

voltage is developed across the load resistor R_1, which has a value of approximately 3300 ohms. In parallel with this load resistor for a-c voltages, is the coupling capacitor C_2 and tuned inductance L_2. L_2 is tuned to resonance at the intermediate frequency and so presents a high impedance in parallel with R_1. A third parallel branch which forms part of the load circuit is provided by two diode valves V_2 and V_3. The i-f output of V_1 is applied to the cathode of V_2 and to the anode of V_3. These two electrodes are biased 0.2 volts negative with respect to ground. The cathode of V_3 is grounded while the anode of V_2 is biased to −0.4 volts with respect to ground.

The output voltage from V_1 varies about the bias voltage. If the amplitude of this voltage is less than the bias voltage, as shown in Fig. 17-43 (a), neither valve conducts. On the positive half cycle, the cathode of V_2 becomes more positive to its anode. At the same time the anode of V_3 becomes less negative to its cathode but does not become positive to its cathode, even at the peak of the cycle. During the negative half of the input cycle, the cathode potential of V_2 approaches the negative potential of the anode but does not cause the valve to conduct. Under these conditions V_2 and V_3 present an impedance of about 30,000 ohms in parallel with the load resistor R_1 and the loading effect on R_1 is negligible. The amplification of the stage is then a maximum.

If the signal voltage exceeds the bias voltage, as shown in Fig. 17-43 (b), each valve conducts during different parts of the cycle. When this occurs, the impedance of the diodes falls to about 330 ohms in parallel with R_1 and L_2. The amplification of the stage therefore falls to a low value for large signals. Five or six stages of logarithmic amplifiers operating at the intermediate frequency are normally used in the receiver. Such a receiver reduces the wave clutter in both type *A* and *PPI* presentations without appreciably cutting down the size of echoes from distant targets. A

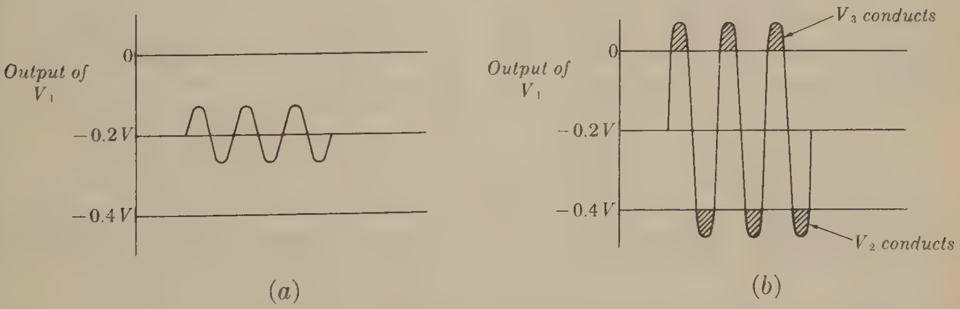


FIG.17-43 Operation of logarithmic amplifier (a) Output from V_1 less than diode bias voltage (b) Output from V_1 greater than diode bias voltage.

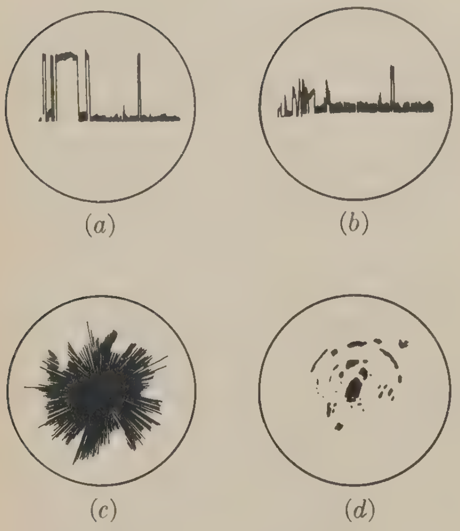


Fig.17-44 Effect of logarithmic receiver on presentation (a) Type *A* scan with linear receiver (b) Type *A* scan with logarithmic receiver (c) *PPI* scan with linear receiver (d) *PPI* scan with logarithmic receiver.

comparison of the appearance of an *A* scan with a linear and with a logarithmic receiver is shown in Fig. 17-44 (a) and (b). The improvement in a *PPI* presentation is shown by the comparison diagrams of Fig. 17-44 (c) and (d).

L. *Automatic Gain Control*

A radar set is sometimes required to follow a target which has been detected by the equipment. This is referred to as *tracking*. It usually requires that the amplitude of the echo to be followed remain constant in amplitude. This is difficult to achieve with a constant-gain receiver as fading ordinarily causes fluctuations in the echo received from a fixed target. *Automatic gain control* (*a.g.c.*) is used to vary the receiver gain in such a manner as to maintain a constant receiver output for a specified target.

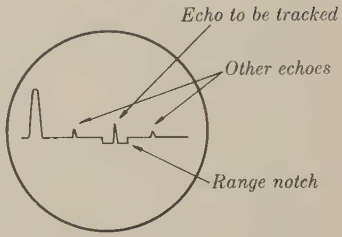


Fig.17-46 Range notch to indicate target selected by *a.g.c.* circuit.

A block diagram of an *a.g.c.* circuit is shown in Fig. 17-45. It operates from the received echo and develops a d-c voltage which increases as the echo signal decreases. This voltage is used as the H.T. supply to one or more amplifying stages in the receiver. It increases the gain of the stage to compensate for the decrease in signal voltage. The first valve stage is a target selector which receives a gate voltage from a delay circuit controlled by the operator of the set. The operator places a *range gate* or *notch* under the echo pulse of the target to be tracked (Fig. 17-46). This range notch and the target selector gate are developed by the same circuit and therefore occur simultaneously. The target selector gate effectively turns on the first stage so that it amplifies in a normal manner during the interval covered by the range notch in the indicator. During the remainder of the sweep the target selector circuit is biased off.

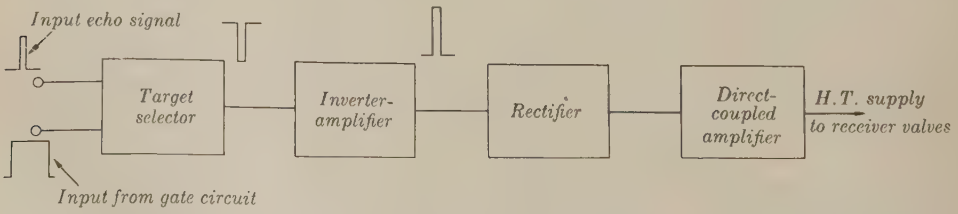


Fig.17-45 Block diagram of *a.g.c.* circuit.

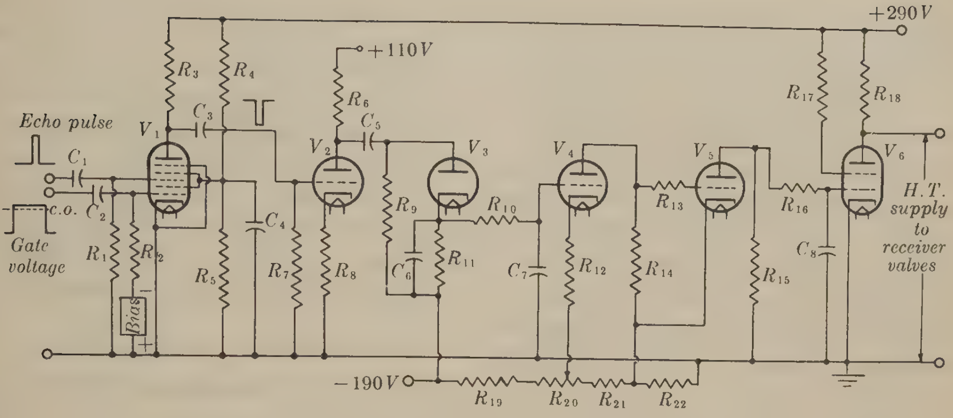

+290 V
+110 V
Echo pulse
C.O.
Gate voltage
Bias
H.T. supply to receiver valves
−190 V
V_1 V_2 V_3 V_4 V_5 V_6
C_1 C_2 C_3 C_4 C_5 C_6 C_7 C_8
R_1 R_2 R_3 R_4 R_5 R_6 R_7 R_8 R_9 R_{10} R_{11} R_{12} R_{13} R_{14} R_{15} R_{16} R_{17} R_{18} R_{19} R_{20} R_{21} R_{22}

FIG. 17-47 Circuit for automatic gain control.

The echo signal voltage from the receiver is also applied to the target selector stage. This valve therefore amplifies only the signal included in the range notch and rejects other signals. The second stage inverts and amplifies this signal and the third stage rectifies it. The output of the rectifier is a d-c voltage which is proportional to the amplitude of the echo signal included in the range notch. This voltage is inverted and amplified by a direct-coupled amplifier and is used as the H.T. supply to the later stages of the receiver. If the amplitude of the received echo increases, the H.T. supply to the receiver decreases and the final output echo pulse remains constant in amplitude.

A complete circuit diagram of the *a.g.c.* circuit is shown in Fig. 17-47. Waveforms of voltages at various points are shown in Fig. 17-48. V_1 is a pentagrid amplifier which is normally cut off by a negative voltage on the first grid. The gate voltage applied to this grid raises the grid potential above cut-off to permit amplification during the period covered by the range notch on the c.r.t. The echo signals are applied to the third grid of V_1. Only that signal which occurs during the range notch appears in the output of the valve. The output is a negative pulse at the anode which is amplified and inverted by V_2. The positive pulses from the anode of V_2 are applied to the diode rectifier V_3. This valve develops a p.d. across its load resistor R_{11} which is proportional to the amplitude of the echo pulse to be tracked. The capacitor C_6 connected in parallel with the load resistor of V_3 maintains

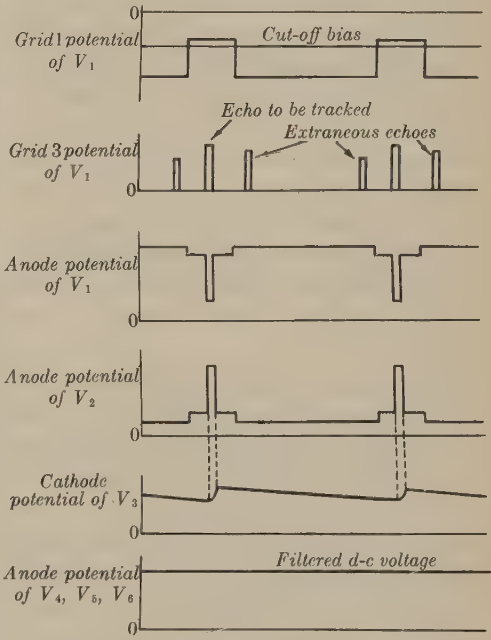


FIG. 17-48 Voltage waveforms developed by *a.g.c.* circuit.

fairly steady voltage across the resistor. This voltage is further filtered by a low-pass filter consisting of a series resistor R_{10} and by-pass capacitor C_7 and applied to the grid of V_4. This valve together with V_5 and V_6 constitute a three-stage direct-coupled amplifier. The anode potential of the final power-amplifier stage is used as the H.T. supply voltage for several receiver valves. Therefore, the amplification of the receiver increases as the anode potential of V_6 rises and vice versa.

Since the anode potential of V_6 must be very positive and since V_4, V_5 and V_6 are all directly coupled a special H.T. arrangement is necessary for V_4 and V_5. The voltage divider used for this purpose consists of R_{19}, R_{20}, R_{21} and R_{22} across a $-190V$ supply. The V_5 anode is at a small negative potential. Hence, the grid of V_6 is at a negative potential. The V_5 cathode is connected to a more negative potential point on the voltage divider. Since the V_5 grid and hence the anode of V_4 must be slightly negative to V_5 cathode therefore the cathode of V_4 must be connected to a more negative point on the voltage divider than the V_4 anode. The grid of V_4 is connected to $-190V$ so that it is more negative than its cathode. This necessitates both the cathode and anode of V_3 being connected to $-190V$. The setting of R_{20} controls the H.T. and bias of V_4 and hence the average value of the d-c output of V_6.

The operation of the circuit is summed up as follows. If the amplitude of the echo signal increases, the positive potential at the cathode of V_3 increases, the d-c potential at the anode of V_4 decreases, the potential at the anode of V_5 increases and the potential at the anode of V_6 decreases. Consequently the amplification of the receiver decreases to compensate for the increased echo signal. The time required for the circuit to respond to changes in amplitude of echo pulses is determined by the time constants of the filter circuits which follow the diode detector. These time constants are so chosen that successive echo pulses can be received with little change in the d-c output from V_6. If, however, there is a change in the average amplitude of the received pulses, the d-c output of V_6 changes to correct the gain of the receiver.

M. Automatic Frequency Control

In describing the operation of the radar circuits in the present chapter, it has been assumed that the frequencies of the transmitter and local oscillator remain constant throughout the operation of the set. This assumption is not justified when dealing with microwave radiations. Both the transmitter and the receiver are subject to changes in operating frequency due to variations in temperature, reflected impedances and other operating conditions. Any change in frequency of the r-f carrier or of the local-oscillator output results in a corresponding shift in the intermediate frequency. If this shift is great enough to bring the i-f outside the band-pass limits of the receiver, the signal is not amplified and the target is lost. The receiver bandwidth may be increased to compensate for frequency drift but this results in an increase in noise level. A better method is to employ automatic frequency control of the local oscillator. The a.f.c. circuit operates from the i-f signal. If this signal varies from the correct frequency, the a.f.c. circuit brings it back to the correct frequency by re-tuning the local oscillator.

An a.f.c. circuit is shown in Fig. 17-49. V_1 is a pentode amplifier which receives the i-f signals from the output of the i-f amplifier in the receiver. The anode load circuit of this amplifier is the transformer T_1. The secondary winding of T_1 is tuned to the intermediate frequency. V_2, T_1 and their associated components

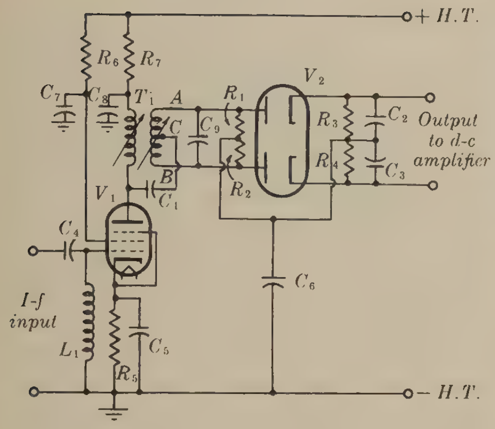


FIG. 17-49 Frequency discriminator for use in a.f.c. circuit.

make up a frequency discriminator circuit of the type used in f-m receivers. The operation of the discriminator is as follows.

The output of V_1 is inductively coupled by T_1 to the anodes of V_2. The tuning of the primary and the coupling between windings are adjusted to produce a voltage in the secondary which is 90° out of phase with the primary voltage. In addition, the output of V_1 is coupled through C_1 to the centre tap of the secondary winding. The voltage between the point A and ground equals the primary voltage coupled through C_1 plus the voltage induced in the top half of the secondary winding by T_1. Similarly, the voltage between point B and ground equals the primary voltage coupled through C_1 plus the voltage induced in the lower half of the secondary winding by T_1. The induced voltages at A and B are 180° out of phase with each other and the voltage coupled through C_1 is intermediate in phase between these two, as shown by the vectors of Fig. 17-50 (a).

The vector E_{OC} represents the voltage between C and ground, which is the output voltage of V_1. The vector E_{CA} represents the voltage induced in the upper half of T_1 secondary. The vector sum of these voltages is represented by E_{OA}. This is the voltage which is applied to the anode of the upper diode of V_2. Similarly, the voltage which is applied to the lower diode is the vector sum of the voltage between C and ground and the voltage induced in CB. The resultant voltage is represented by the vector E_{OB}.

When the i-f signal is at the correct frequency, E_{CA} leads E_{OC} by 90°, and E_{CB} lags E_{OC} by 90°. As a result, the magnitudes of E_{OA} and E_{OB} are equal. That is, the anode of the upper diode of V_2 is driven the same amount positive during one half cycle as the lower anode is made positive during the other half cycle. The fact that the anode voltages are out of phase with each other does not affect the amount of current which each valve conducts.

When the upper anode is made positive, electrons flow from cathode to

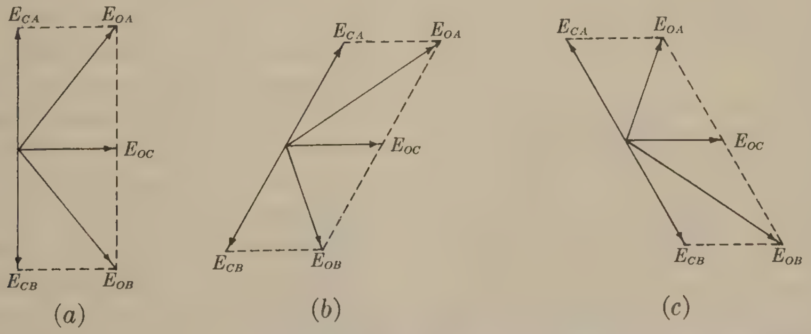


FIG. 17-50 Vector diagrams of input voltage to V_2 when (a) i-f signal is at the correct frequency (b) i-f signal is below the correct frequency (c) i-f signal is above the correct frequency.

anode, down through R_1 and up through R_3. This develops a p.d. across R_3 making the top end positive to the bottom end. When the lower anode is made positive, electrons flow from cathode to anode, up through R_2 and down through R_4. The p.d. across R_4 makes the lower end positive to the upper end. Capacitors C_2 and C_3 maintain the p.d.'s across R_3 and R_4 steady during the period between i-f signal pulses. Since the p.d.'s across R_3 and R_4 are equal and of opposite polarity, the d-c voltage across the output terminals of the discriminator is zero.

When the i-f signal is below the correct frequency, the voltage across the upper half of the T_1 secondary leads the primary voltage by less than 90°. The voltage across the lower half of the secondary winding lags the primary voltage by more than 90°. The vectors representing these voltages take the positions shown in Fig. 17-50 (b). The magnitude of E_{OA} is now greater than the magnitude of E_{OB}. Consequently, the upper anode of V_2 is driven more positive than the lower anode, and therefore the upper diode conducts more current. The p.d. developed across R_3 is therefore greater than that across R_4. The d-c output voltage is therefore positive at the upper terminal. This d-c voltage is amplified by a d-c amplifier and fed back to the local-oscillator circuit. In the case of a reflex klystron, the voltage is usually applied to the repeller grid to change the frequency of the klystron in such a way as to bring the i-f back to the correct frequency.

When the i-f signal is above the correct frequency, the phase relations of the voltages in the secondary of T_1, are such as to make E_{OB} greater than E_{OA} as shown in Fig. 17-50 (c). Therefore, the lower diode conducts more current than the upper diode. The p.d. across R_4 is greater than that across R_3 and a negative output is developed by the discriminator. This voltage, amplified by the d-c amplifier again causes the klystron to change frequency in such a manner as to correct the i-f.

The a.f.c. circuit not only corrects frequency drift of the local oscillator, but also maintains the correct i-f when the transmitter frequency shifts. This makes it possible for the receiver to amplify echo pulses that would otherwise be lost.

17.5 T-R and anti-T-R devices. The purpose of the aerial system of a radar set is to conduct r-f energy from the transmitter to the radiating elements, radiate it in a directional beam, receive the returning echo and conduct it back to the receiver. If the aerial system is to be an efficient one, a minimum of losses must occur in the transmission lines or waveguides which make up the system, and in the transmitting and receiving aerials.

When a radar receiver and transmitter are operated close to each other, it is usually difficult to prevent the strong transmitter pulse from overloading the receiver. In microwave radar sets, the first stage in the receiver chain is a mixer unit which contains a crystal first detector. The crystal must be protected from overload caused by the strong transmitter pulse. Even in equipment in which a crystal is not employed, the receiver must be protected from strong transmitter pulses which would cause blocking in the *R-C* coupling circuits. If blocking is allowed to occur, the receiver becomes paralyzed and echo pulses are lost until the p.d. across the coupling capacitor is restored to its normal working value.

The receiver may be protected to some extent by using separate transmitting and receiving aerials mounted one above the other to prevent directing transmitter pulses into the receiving aerial. There still remains the danger of overload due

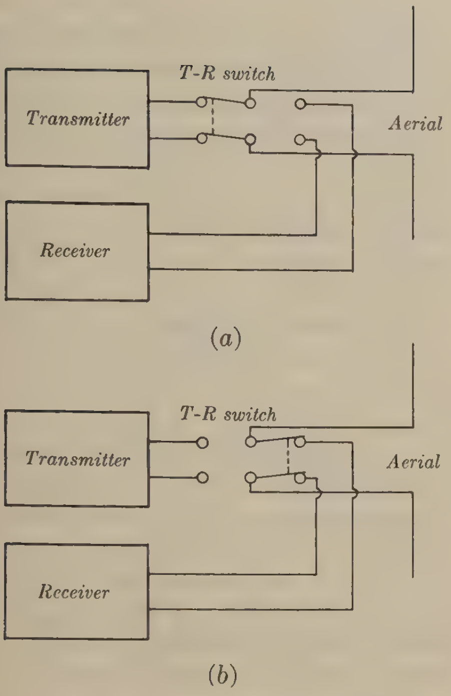


FIG.17-51 Function of T-R switch (a) while transmitter is pulsing (b) while transmitter is not pulsing.

to reflection of the transmitter pulse from nearby objects. In addition, the receiver may be turned off for the duration of the transmitter pulse by a gate pulse applied to the receiver chain. This, however, does not protect the receiver of one radar set from strong pulses from another radar transmitter nearby operating independently on the same frequency.

A better method of protecting the receiver is by the use of a *transmit-receive (T-R) switch.* This device is usually employed in microwave radar equipments in which a single aerial is used for transmitting and receiving. The *T-R* switch operates as a double-pole double-throw switch (Fig. 17-51) which effectively connects the aerial to the transmitter during the transmitting pulse and connects the aerial to the receiver at other times.

A mechanical switch is impractical for such an application because the duration of the transmitter pulse is seldom greater than two microseconds and the switching action must be repeated several hundred times per second. Consequently an electronic switch is used.

Not only must the transmitter be prevented from pulsing into the receiver chain, but the returning echo pulse must be prevented from being absorbed by the transmitter if the aerial system is to be efficient. In other words, the impedance of the aerial must be matched to that of the transmitter during the radiation period, and matched to the receiver during the remainder of the operating cycle. The transmitter must be mismatched to the aerial during the reception period.

A. *Spark-Gap T-R Switch*

A method of accomplishing this switching in an open two-wire transmission line is illustrated in Fig. 17-52. The junction of the aerial, transmitter and receiver lines is called the *T-junction.* In order to simplify the problem, it is assumed that the characteristic impedance of the transmission line, the impedance of the transmitter when transmitting, the impedance of the aerial and the impedance of the receiver are all equal. If this is not so, the impedances may be matched by suitable transformers. It is also assumed that the impedance of the transmitter rises appreciably when

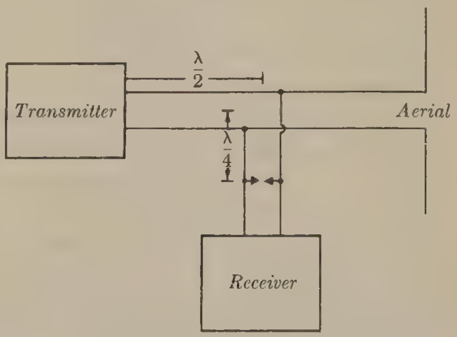


FIG.17-52 Spark-gap T-R switch.

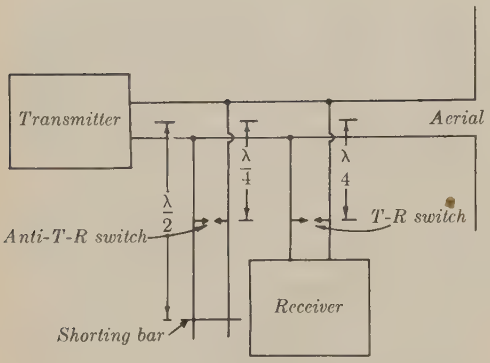


Fig. 17-53 Anti-T-R switch in conjunction with T-R switch.

the transmitter is not pulsing. The *T-R* switch has a spark gap which is placed between the two conductors of the receiver line one-quarter wavelength distant from the T-junction. The gap is adjusted so that it arcs across when a high voltage is presented to it. For low voltages the spark gap does not arc.

When the transmitter pulses, the high voltage developed between the conductors of the transmission line causes the gap to arc and present a low impedance at that point. This impedance is considerably lower than that of the line. At the T-junction, therefore, a quarter wavelength distant from the spark gap, a high impedance is presented to the transmitter pulse in the direction of the receiver. The low impedance of the transmitter is matched to the low impedance of the aerial. The r-f energy from the transmitter therefore flows to the aerial and is largely prevented from entering the receiver.

When an echo pulse enters the aerial, it produces a voltage across the conductors of the transmission line which is not sufficient to break down the resistance of the spark gap. At this time the spark gap has no effect on the receiver branch. The low impedance of the aerial is matched to the low impedance of the receiver for maximum transfer of energy. At the same time, the impedance of the transmitter is large. This high impedance is presented at the T-junction by placing the transmitter one half wavelength (or a multiple of half wavelengths) from the T-junction. In this way the transmitter is prevented from absorbing energy from the echo pulses. If the impedance of the transmitter falls at the end of the pulse, a quarter-wave line is required between the T-junction and the transmitter in order to present a high impedance at the T-junction.

B. *Anti-T-R Switch*

The impedance of the transmitter does not always change sufficiently at the end of the pulse to permit the use of resonant transmission lines to block received pulses from the transmitter. In this case, a second spark gap called the *anti-T-R spark gap* is required (Fig. 17-53). The anti-T-R spark gap is also connected in a section of transmission line which forms a T-junction with the line from the transmitter to the aerial. This T-junction is placed one quarter wavelength nearer the transmitter than the T-junction to the receiver. The anti-T-R gap is placed one quarter wavelength from its T-junction. The line containing the anti-T-R gap is one half wavelength long and is terminated at the far end by a short circuit between the two conductors. The operation of the anti-T-R switch is as follows.

When the transmitter pulses, the anti-T-R gap arcs and becomes a low impedance. This is transformed to a high impedance and the anti-T-R switch absorbs very little of the transmitter pulse, just enough to maintain the arc. When the echo pulse is received, the anti-T-R gap does not arc and therefore acts as an open circuit. The shorted end of the line presents a short circuit one half wavelength distant, at the T-junction. Therefore, one quarter wavelength along

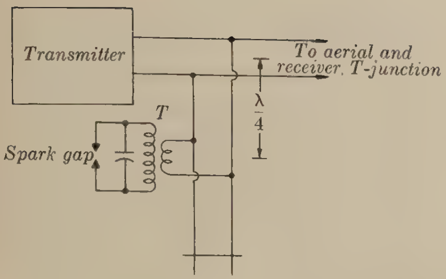


FIG.17-54 Resonant transformer to increase efficiency of anti-T-R switch.

the main transmission line, a high impedance is presented in the direction of the transmitter. The impedance of the transmitter is of no consequence in this case as it is in parallel with the short circuit presented at the T-junction by the shorted half-wave line.

C. Methods of Improving the Efficiency of Spark-Gap Switches

As pointed out above, the spark gap requires some energy from the transmitter to maintain the arc during transmission. This reduces the amount of energy passed from the transmitter to the aerial and lowers the efficiency of the system. In order to reduce the energy lost and to present a higher impedance at the T-junction it is necessary to make the T-R gap present a lower impedance when conducting. This may be accomplished by using a step-up transformer between the point of the T-R line to be shorted and the actual T-R switch. This arrangement is shown in Fig. 17-54. A transformation ratio of about 1:10 is used to increase the efficiency without causing the switch to arc during the reception of echo pulses. By using the transformer, a smaller voltage is required in the transmission line to start and maintain the arc in the spark gap. In addition, the impedance reflected into the transmission line is reduced to one-tenth of the impedance of the gap while conducting. This in turn reflects an impedance at the T-junction equal to 10 times the impedance presented without the transformer.

While the transformer of Fig. 17-54 is shown as a coil-type r-f unit, a resonant line is actually used because of its greater efficiency at radar frequencies. A typical quarter-wave resonant-line transformer to increase the efficiency of the anti-T-R switch is shown in Fig. 17-55. While the open-wire line shown in the diagram is suitable for use in the 500 Mc/s range, an appreciation of its operation is fundamental to the understanding of waveguide transformers for switches at microwave frequencies.

The transformer consists of a quarter-wave section of transmission line connected to the anti-T-R branch one quarter

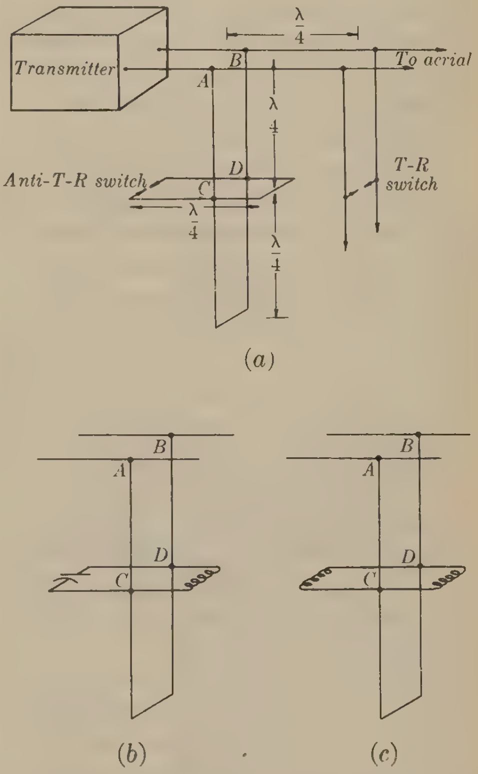


FIG.17-55 Quarter-wave resonant-line transformer for anti-T-R switch (a) Transformer and spark gap (b) Equivalent circuit of transformer when gap does not arc (c) Equivalent circuit of transformer when gap arcs.

wavelength from the T-junction. A short circuit is placed across the end of the shorter section and the anti-T-R spark gap across the end of the longer section of the transformer. The equivalent circuit of the transformer when the spark gap is not conducting is shown in Fig. 17-55 (b). The shorted end of the transformer is less than a quarter wavelength long and presents an inductive reactance. The open end of the transformer is also less than a quarter wavelength and therefore presents a capacitive reactance. These reactances are equal so that the transformer acts like a parallel-resonant circuit of high impedance at the operating frequency. A low impedance is therefore reflected at the anti-T-R junction and a high impedance at the T-R junction. Hence the echo pulse is prevented from following the transmission line leading to the transmitter.

The equivalent circuit of the transformer during transmission is shown in Fig. 17-55 (c). The gap is made to arc and acts as a low impedance. Therefore the gap end of the transformer is a shorted section less than a quarter wavelength long and presents an inductive reactance. This reactance is in parallel with the small reactance of the shorted end of the transformer. The two reactances in parallel present a low impedance across the line at the point of connection to the transformer. Therefore a high impedance is placed across the main line at the T-junction so that only a small amount of transmitter energy is lost in the anti-T-R line.

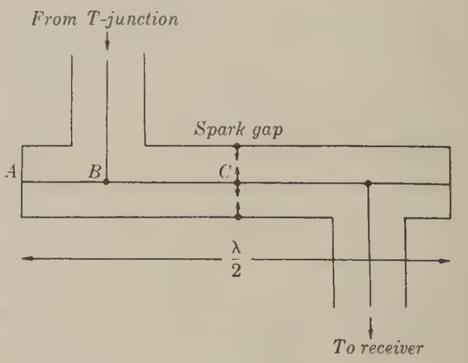


Fig.17-56 Half-wave T-R transformer for use with coaxial transmission line.

The use of T-R and anti-T-R transformers is especially essential when coaxial transmission lines are used in the aerial system. In this case, the impedance of the line is comparable with the impedance of the spark gap when conducting. The impedance of the gap must be transformed to a much lower impedance in order to reflect a high impedance at the T-junction. A transformer for use in the T-R branch of a coaxial transmission line is shown in Fig. 17-56. It consists of a half-wave section shorted at both ends with the spark gap in the centre. When the gap is non-conducting, there is a high impedance between inner and outer conductors at the centre of the transformer and a low impedance at both ends. The lines from the T-junction and the receiver are tapped into the transformer at points which provide a suitable match when the gap is in a non-conducting condition. Therefore energy can pass from the T-junction through the transformer to the receiver with little attenuation. When the transmitter is pulsed, the gap arcs presenting a low impedance. The line from the T-junction is then terminated in two shorted stubs *AB* and *BC* in parallel. Each stub is less than a quarter wavelength and presents a low inductive impedance at the point where the line joins the transformer. The line from the T-junction is then resonant and is adjusted in length to reflect a high impedance at the T-junction in the direction of the receiver. The transformer and T-R switch therefore protect the receiver from damage by the transmitter pulse.

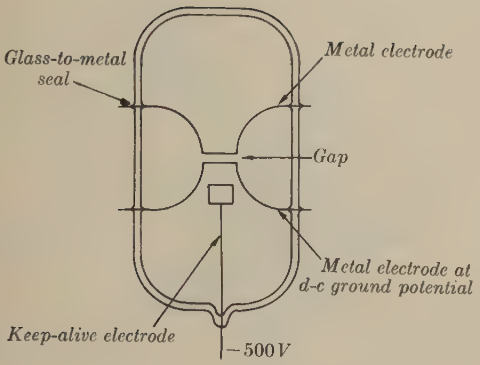


FIG. 17-57 Sealed T-R spark gap.

D. Sealed T-R Spark Gap

While the spark-gap terminals may be exposed to air at atmospheric pressure when used with open-wire lines and coaxial cables at low radar frequencies, they do not meet the requirements necessary for best operation. These requirements are that the resistance of the gap be very high until arcing takes place, and then fall to a low value throughout the arc. At the end of the arcing period, the resistance should rise very quickly to its original value. A spark gap in air ordinarily has an arcing resistance in the neighbourhood of 50 ohms and requires about 10 microseconds for de-ionization of the air to take place and extinguish the arc. Both the resistance and the de-ionization period can be reduced by placing the electrodes inside a glass envelope and reducing the pressure of the enclosed gas. The de-ionization time can be reduced to about one microsecond by placing water vapour in the tube at a pressure of one millimeter of mercury.

The operation of the switch is further improved by introducing a third electrode into the glass envelope as shown in Fig. 17-57. This electrode, known as the *keep-alive* is placed close to one terminal of the gap and maintained at a negative potential of 500-1000 volts. A small discharge is maintained between the keep-alive and the neighbouring electrode, but the ionization does not extend to the main gap until the transmitter pulse enters the switch. The effect of the keep-alive electrode is therefore to reduce the time required for the arc to be established across the gap and also to prevent stray ions in the switch from reaching the main gap and producing noise in the receiver.

Most of the T-R switches used at microwave frequencies are fitted with a high-Q metal resonant switch similar in construction to the reflex klystron but with only the keep-alive electrode instead of the filament, cathode, grids and repeller (Fig. 17-58). Coupling loops are provided for inserting the r-f energy from the aerial into the cavity and for extracting the r-f energy from the cavity for the receiver. The cavity is tuned to resonate at the transmitter frequency by adjusting the tuning plugs as in the case of the klystron. A sectional view of a T-R switch connected in a coaxial transmission line is shown in Fig. 17-59.

While T-R switches are comparatively simple in construction, they tend to de-

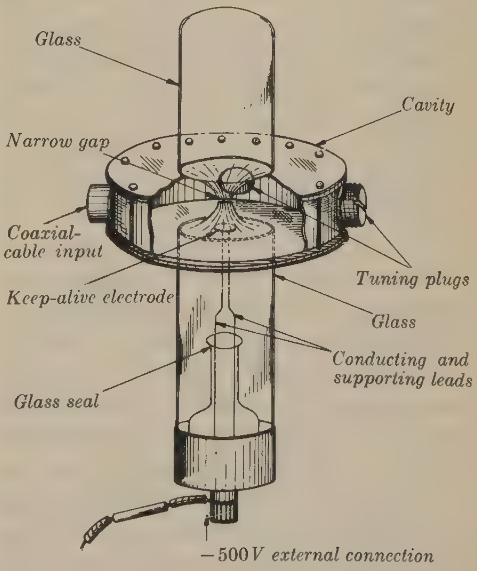


FIG. 17-58 Cavity T-R switch.

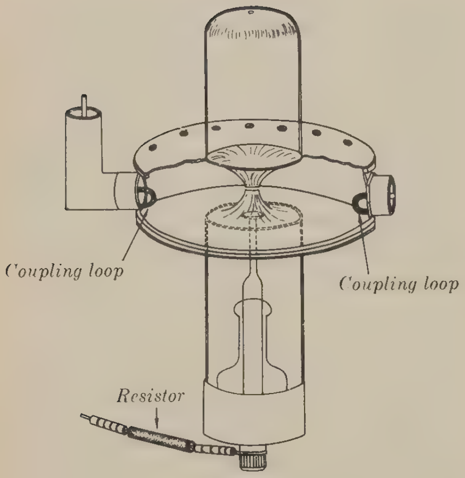


FIG. 17-59 Section of T-R switch connected in coaxial transmission line.

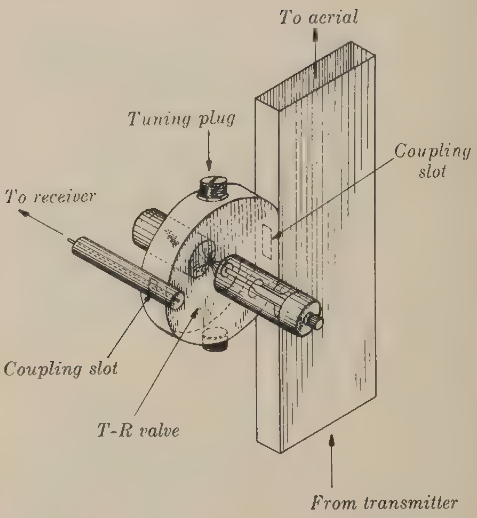


FIG. 17-60 T-R cavity directly coupled to waveguide.

teriorate with continued use and should be checked periodically to ensure optimum operation. Metal particles dislodged from the electrodes tend to form a conducting layer on the glass walls and reduce the Q of the cavity. This increases the losses in the cavity during the receiving period and results in reduced input to the receiver. Another cause of deterioration is the absorption of gas within the envelope by the metal electrodes. The switch then requires a higher voltage to establish the arc and consequently it affords less protection to the receiver.

E. T-R Switches for Waveguide Systems

Resonant-cavity switches may be used in waveguide transmission systems by employing direct or indirect coupling to the waveguide. If the indirect method is used, a section of coaxial line is connected between the waveguide and the resonant cavity. Energy is extracted from the waveguide by means of a probe and inserted into the cavity by a coupling loop. In this case the operation of the T-R device is the same as if a coaxial line is used throughout.

The T-R cavity may be directly coupled to the waveguide as shown in Fig. 17-60. Coupling is provided by a coupling slot or aperture which is common to the cavity and the waveguide. The r-f energy from the transmitter passes along the waveguide, through the aperture and excites the cavity causing ionization to take place in the spark gap. This de-tunes the cavity to prevent energy from passing through to the receiver. It also presents a low impedance across the aperture which effectively seals the opening and permits the transmitter energy to flow to the aerial. When an echo is received, the cavity is also excited but not sufficiently to cause the spark gap to arc. The high Q of the cavity is maintained and the energy is fed through to the receiver coupling slot on the opposite side.

Instead of coupling the T-R and anti-T-R switches directly to the main waveguide, these devices may be placed at the ends of branch waveguides as shown in Fig. 17-61. Each branch guide is made one half wavelength long (or a multiple of half wavelengths) so that the

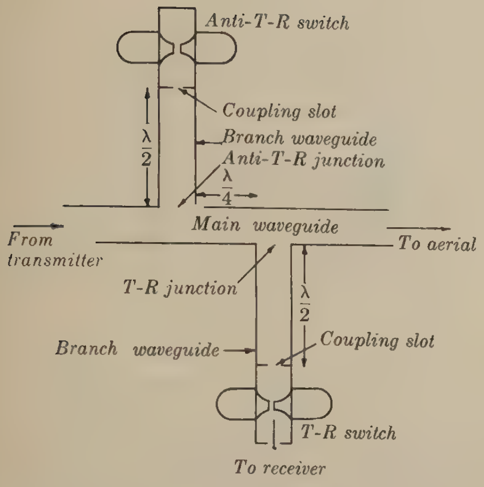


FIG.17-61 T-R and anti-T-R switches in branch waveguides.

impedance it presents at the main guide is the same as the impedance at the end remote from the main guide. The remote end is terminated in a coupling slot which is common to the waveguide and the resonant cavity of the switch.

When the transmitter is pulsed, sufficient energy flows into the branch waveguides to excite the resonant cavities and cause the spark gaps to arc. This detunes the cavities and produces a short-circuiting effect across each of the coupling slots. Since each guide is a half wavelength long, this short circuit also appears across the T-junctions to the main guide. Hence the transmitter energy flows to the aerial.

During the period between transmitter pulses when echoes are received, neither cavity is excited sufficiently to cause arcing at the gap. Energy is therefore permitted to flow through the coupling slots with little attenuation. Pulses from the aerial flow to the T-R junction, into the half-wave section of guide and through the high-Q resonant circuit to the receiver. The anti-T-R switch reflects an open circuit into the branch guide half a wavelength away. This impedance is transformed by the quarter-wave section of the main guide between the T-junctions to short the opening of the guide in the direction of the transmitter at the T-R junction. Thus the echo pulse is prevented from following the main guide back to the transmitter and is diverted into the receiver guide.

F. Echo Boxes

The tuning of the i-f amplifiers in the receiver of a radar set is usually pre-set and not adjusted during operation. The tuning of the transmitter is also fixed in the case of microwave radar. The only frequency adjustment made during operation is that of the local oscillator. It must be kept operating at a frequency which differs from the transmitter frequency by the i-f. This adjustment is loosely referred to as tuning the receiver. In addition there are ordinarily several adjustments in the aerial system which depend upon the frequency of the transmitter and which must be set to obtain maximum efficiency of the set. These include the adjustment of tuning stubs used for impedance matching, T-R switches, anti-T-R switches, etc.

The most satisfactory method of setting these controls is by obtaining radar echoes from a distant target. The adjustments are made to give maximum amplitude of echoes on the c.r.t.

In the case of shipborne or airborne radar, echoes may not always be available. In this case an echo box is useful for tuning the set.

The echo box consists of a resonant cavity with a very high Q, of the order of 35,000. The cavity is tuned to the transmitter frequency and is coupled to the transmitter aerial to receive energy when the transmitter pulses. Oscillations set up in the cavity continue for a number of microseconds after the transmitter pulse has ended. This oscillation of the cavity is called *ringing*. While the ringing continues, the cavity feeds energy

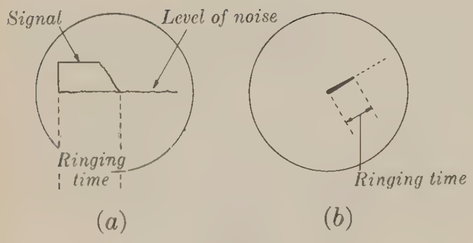


Fig. 17-62 Tuning the receiver by means of an echo box (a) Type *A* presentation (b) *PPI* presentation.

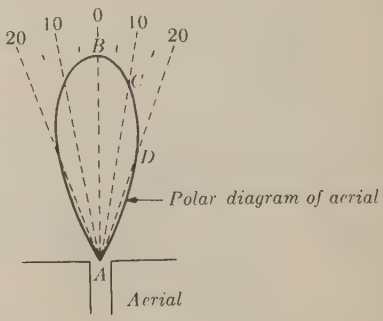


Fig. 17-63 Polar diagram of a directional aerial.

into the receiver at the frequency of the transmitter. The effect is the same as if echoes were returning from nearby targets. The aerial system and local oscillator can then be tuned to produce the longest signal on the c.r.t. as shown in Fig. 17-62 (a). The ringing time is considered to end when the signal falls to the level of the noise on the screen.

The echo box may be excited by placing it in front of the transmitting aerial. It is usually preferable however to excite the cavity by placing a small pick-up dipole in front of the aerial and feeding the echo box by a coaxial line. In this way, the echo box may be placed close to the indicator, if desired.

The echo box is tuned to the transmitter frequency by adjusting its tuning plugs to give maximum ringing time on the c.r.t. The tuning is then kept constant while the receiver is tuned. When this has been accomplished, the echo box should be thoroughly de-tuned so that it will not absorb energy during operation of the set.

The echo box method of tuning is preferable to tuning on permanent echoes if only a *PPI* is available as the indicator. If a *PPI* is used, tuning on a permanent echo requires that the controls be adjusted for maximum intensity of the echo spot on the screen. This may be difficult to judge because *PPI* tubes usually have long persistence screens. If an echo box is used, the controls are adjusted to give the maximum length of bright trace from the centre of the screen as shown in Fig. 17-62 (b).

17.6 Lobe switching. The accuracy with which the bearing of a target may be determined depends upon the directional properties of the transmitting and receiving aerials. An indication of the directional properties of the aerial is provided by the polar diagram as shown in Fig. 17-63. The diagram indicates the relative intensity of the field radiated in various directions from the aerial, when the aerial is used for transmitting purposes. For example, the lengths *AB* and *AC* represent the relative intensities of radiation in the direction at right angles to the dipole and 10 degrees away from this direction. In the case of the aerial being used for reception the polar diagram indicates the relative input from the aerial to the receiver for signals received from various directions.

When a radar bearing of a target is being determined the aerial is rotated until the maximum input signal to the receiver is produced as indicated by the maximum amplitude of the echo on the c.r.t. The aerial is then pointing in the direction of the target. This method, however, does not make use of the maximum sensitivity of the aerial to changes in bearing as indicated by the polar diagrams of Fig. 17-64. A 10-degree shift of the aerial about the position of maximum sensitivity, Fig. 17-64 (a), pro-

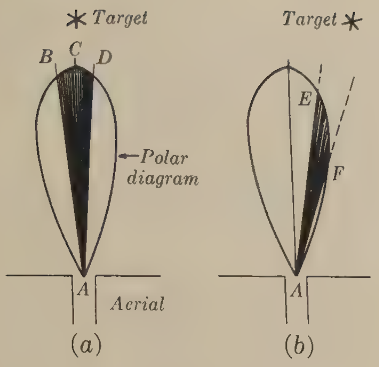


FIG.17-64 Effect of bearing of target on sensitivity (a) Maximum sensitivity, minimum sensitivity to changes in bearing (b) Reduced sensitivity, greater sensitivity to changes in bearing.

duces a very small variation in sensitivity as indicated by the slight differences in the lengths *AB*, *AC*, *AD*. If, however, the target is 15 degrees off the position of maximum signal, Fig. 17-64 (b), a 10-degree shift of the aerial position results in a much greater variation in signal input as indicated by the relative lengths of *AE* and *AF*. In order to make use of the region of maximum sensitivity to changes of bearing two aerials may be used with these regions overlapping as in Fig. 17-65. Each aerial provides an input signal to a separate receiver and hence to the c.r.t. where the signal amplitudes are compared; or, each aerial is connected alternately to one receiver and hence to the indicator. This process is called *lobe switching*. The c.r.t. is rotated 90° from its normal position for type *A* presentation so that the trace is vertical and the sweep takes place from the bottom to the top of the screen. The video outputs from the receiver or receivers are applied to the horizontal-deflection plates to deflect the sweep sideways from its centre position.

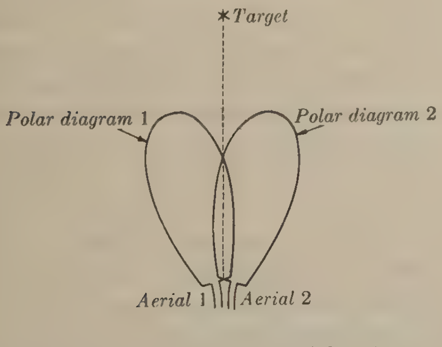


FIG.17-65 Polar diagram of aerial system used for lobe switching.

Fig. 17-66 shows an arrangement using two receivers. The two receivers are alternately connected to the c.r.t. by a double-pole, double-throw switch which is operated by a switch motor. The switching takes place many times per second so that the two traces appear simultaneously on the c.r.t. The sequence of events is, therefore, as follows. The switching motor connects aerial 1 and receiver 1 to the c.r.t. The transmitter is pulsed and the sweep takes place from the bottom to the top of the screen. The transmitter pulse which appears at the bottom of the trace and any echoes received during the sweep are made to deflect the trace to the left. When this cycle has been repeated several times, the switching motor disconnects receiver 1 from the c.r.t. and connects receiver 2 in its place. The output of this receiver is of the correct polarity to deflect the

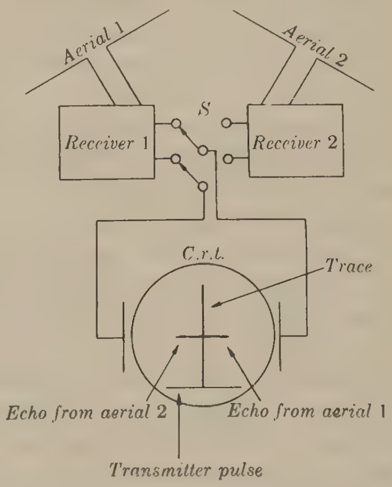


FIG.17-66 Block diagram of lobe-switching arrangement using two receivers.

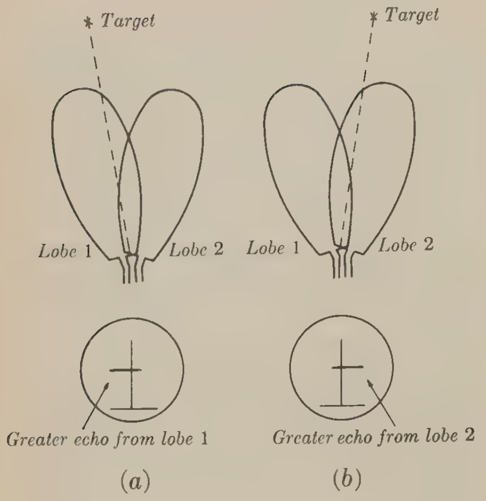


FIG.17-67 Effect of rotating aerial with lobe switching when (a) target is to left of aerial-system centre line (b) target is to right of aerial-system centre line.

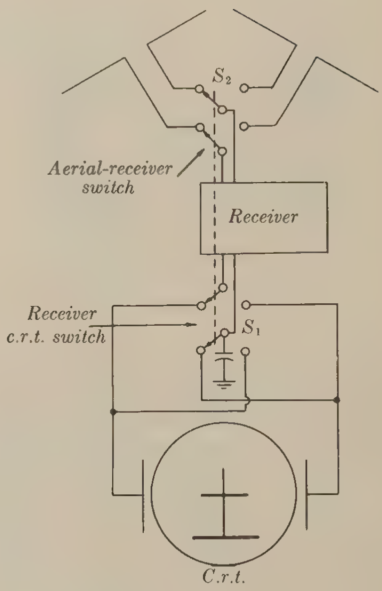


FIG.17-68 Lobe switching with a single receiver.

trace to the right. The echo pulses must appear directly opposite each other along the trace because the position of each indicates the range of the target and is independent of the lobe switching.

When the target is symmetrically placed between the lobes of the two aerials, as shown in Fig. 17-65, the input signals to the two receivers are equal. The receiver gains are then adjusted to produce equal outputs to the c.r.t. so that the echo pulses have the same amplitude. This balance must be maintained during operation.

If the aerial system is rotated so that aerial 1 is pointing more directly at the target than aerial 2, Fig. 17-67 (a), the output of receiver 1 is greater than the output of receiver 2 and the echo pulse on the left side of the trace is greater than that on the right. Similarly, if the line to the target lies more in the polar diagram of aerial 2 than in that of aerial 1, Fig. 17-67 (b), the echo pulse is greater on the right side of the trace than on the left.

In operation, the aerial system is rotated to such a position that the right and left hand echoes are equal. The aerial system is then pointing in the direction of the target. The sensitivity of the lobe-switching system to changes of bearing is greater than that of a single aerial because each aerial is operating over a region of its polar diagram which is more sensitive to changes of bearing. The operator is able to select the correct on-target position of the aerial more readily since he is comparing two signals having greater changes in amplitude per degree change in aerial direction and having opposite directions of change.

While the system shown in the block diagram of Fig. 17-66 is easily understood, it is not the most efficient arrangement because two receivers are required. The same receiver may be used to amplify signals from both aerials by employing a second switch S_2 as shown in Fig. 17-68. This switch is ganged with the c.r.t. switch S_1 and operated by the switch motor. When the switches are in the left hand position, aerial 1 is connected to the input of the receiver. The output

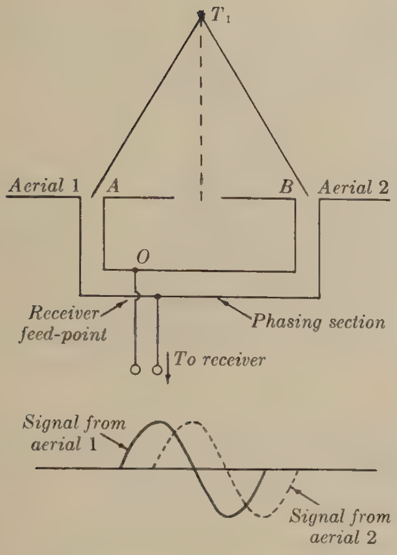


FIG.17-69 Off-centre feed, target on centre line of aerial system.

signal is applied to the left deflection plate while the right hand plate is grounded for a-c. When the switch is thrown to the right the receiver is connected to aerial 2 and the output is applied to the right hand plate of the c.r.t.

The system of using two aerials alternately to effect lobe switching is simple but inefficient because only one aerial is in actual use at any one time. It is possible to make use of the signals from both aerials continuously by shifting the phase of one input to the receiver relative to the other. This has the effect of producing a polar diagram which is shifted from the axis of symmetry of the two aerials. Several methods which have been devised to produce this effect are described in the following paragraphs.

A. *Off-Centre Feed to Receiver*

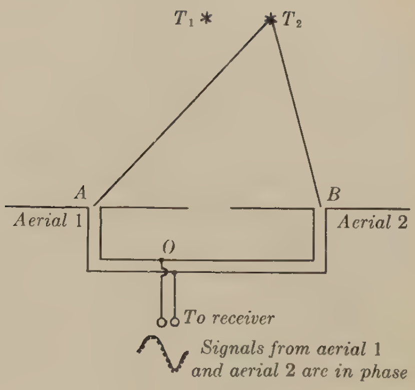


FIG.17-70 Off-centre feed, target in position for in-phase signals to receiver.

A receiving aerial consisting of two dipoles placed end to end is shown in Fig. 17-69. A phasing section is connected to the centre of each dipole as in a balanced collinear aerial. Instead of connecting the receiver feed line to the centre of the phasing section, it is displaced somewhat as shown in the diagram. If a target lies at T_1 on the centre line through the aerials, the echo signal induces voltages at the centre of each dipole which are equal in amplitude and in the same phase. These signals are made to travel unequal distances *AO* and *BO* from the dipoles to the receiver feed point. This causes the two signals to be out of phase with each other as shown.

If, now, the target is moved to the right as shown in Fig. 17-70, the voltage induced in aerial 1 by the echo lags the voltage induced in aerial 2 because of the additional distance from the target to aerial 1. These voltages are brought in phase at *O* because the leading voltage from aerial 2 travels a greater distance in passing from *B* to *O* than the voltage from aerial 1 travels in passing from *A* to *O*. As a result, the input to the receiver has a greater amplitude when the target is at T_2 than when the target is at T_1. The polar diagram of this system is shown in Fig. 17-71.

If the receiver feed point is moved to the right of centre on the phasing section (Fig. 17-72), the polar diagram of the aerial is shifted to the left. By switching from feed point *A* to feed point *B* the

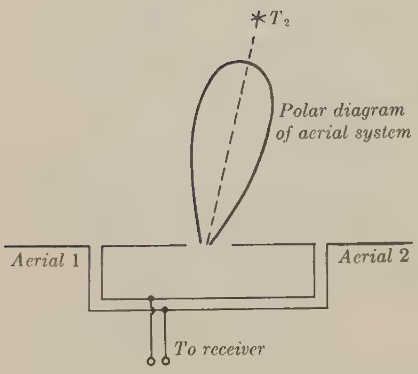


FIG.17-71 Polar diagram for aerial system with off-centre feed to receiver.

same type of indication is produced as by switching from one slanted aerial to another as in Fig. 17-68. That is, when the aerial system is pointed directly at the target the right and left hand echoes have the same amplitude. This arrangement has the advantage that both dipoles provide input signals to the receiver at all times. However, the problem of switching from one feed point to the other is difficult to solve mechanically.

B. Phasing Section Connected to Aerial Elements

The phasing section which effects the lobe switching may be connected directly to the aerial elements (Fig. 17-73) instead of in the transmission lines. The feed lines to the receiver are then connected to the centres of the dipole elements and are switched to the receiver input alternately as in the case of off-centre feed aerials.

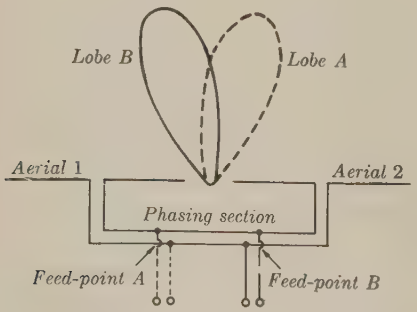


FIG.17-72 Complete polar diagram for aerial system with off-centre feed to receiver.

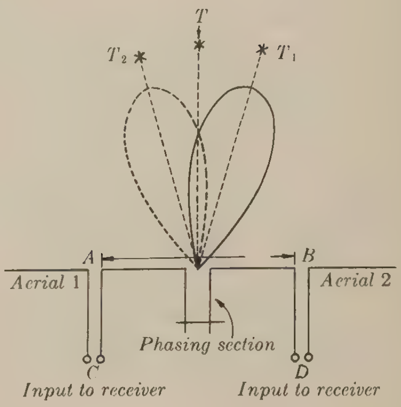


FIG.17-73 Phasing section connected to aerial elements.

When the target is on the centre line perpendicular to the aerials at position T, equal voltages in phase with each other are induced at A and B by the echo. If the receiver is switched to the input at C, one signal travels the path AC to the receiver feedpoint and the other BAC to the same point. These signals may be in or out of phase with each other depending upon the path difference from B to A, including the phasing section. The path difference is adjusted by varying the length of the phasing section so that these voltages are somewhat out of phase with each other and the input to the receiver is less than a maximum. Similarly, if the receiver is switched to D the path difference is the same and the input to the receiver is less than a maximum. If the target is moved to the right to position T_1 as shown in Fig. 17-73, the voltages induced at the centres of the two dipoles are out of phase with each other because of the difference in their distances from the target. The voltage at A must lag behind the voltage at B. The phasing section is now adjusted so that the delay

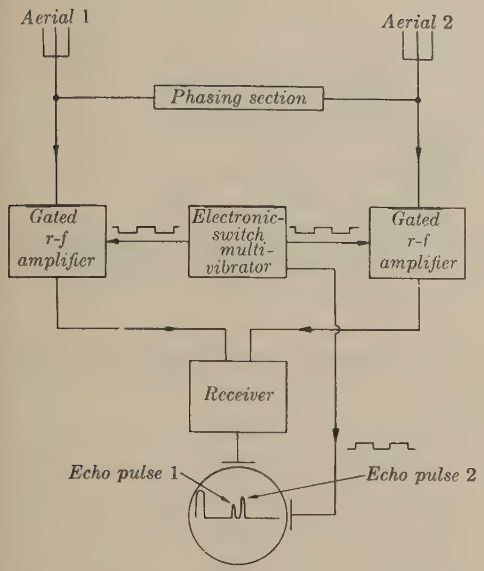


FIG.17-74 Electronic switching of aerials to receiver.

introduced by the path difference from *B* to *A* brings the two signals in phase again when the input to the receiver is at *C*. The input to the receiver is then a maximum and the polar diagram for the input at *C* is displaced in the direction of the target as shown in the figure.

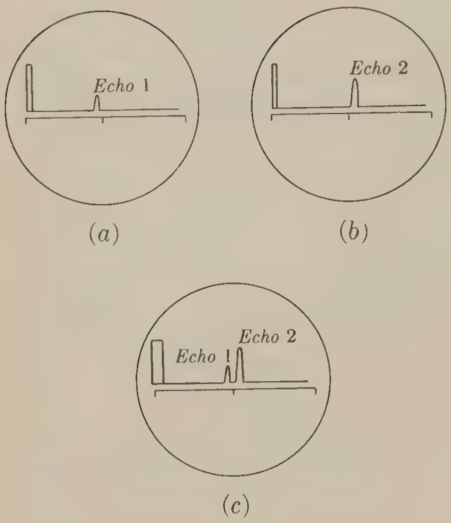
If the target is moved a corresponding amount to the left, to position T_2, the two signals are in phase when the input to the receiver is at *D*. The polar diagram for the input at *D*, as shown by the dotted line, is displaced in the direction of the target as shown.

C. Electronic Switching of Aerial Signals to Receiver

The switching arrangement shown in Fig. 17-68 is subject to mechanical and electrical failure at the switch contacts. The moving parts may be eliminated by employing an electronic switch to connect the aerials to the receiver alternately. A block diagram of this arrangement is shown in Fig. 17-74.

The aerial signals are fed through separate r-f amplifier circuits. These amplifiers are alternately gated on and off by the electronic-switch multivibrator. This circuit produces a square wave of voltage at a frequency which is a submultiple of the pulse repetition frequency of the set. Two voltages 180° out of phase with each other are applied to the screen grids of the gated amplifiers. When the screen of one valve is made positive for normal amplification, the screen of the other valve is made sufficiently negative to cut off the anode current. The signal from aerial 1 is therefore permitted to pass to the receiver while that from aerial 2 is blocked. On the other half cycle of the multivibrator output, the *On* valve is turned off and the *Off* valve is turned on. Therefore the signal from aerial 2 is passed to the receiver while that from aerial 1 is blocked.

When electronic switching of the aerial inputs is employed, the presentation of the echo pulses on the c.r.t. is somewhat different from that previously described. The trace is horizontal as in a normal type *A* presentation but the echo signals from the two aerials are placed side by side as in Fig. 17-74. This is accomplished by applying the square wave of voltage from the electronic-switch multivibrator to the horizontal deflection plates, in addition to the sweep voltage. The square wave causes the trace to shift abruptly to the right during one half of the multivibrator cycle and to the left during the other half cycle. The trace-shifting voltage is so phased that when aerial 1 is connected to the receiver, the trace is shifted to the left, Fig. 17-75 (a), and when aerial 2 is connected to the receiver the trace is shifted to the right, Fig. 17-75 (b). As a result the echo signals from the two aerials appear side by side on the c.r.t. as shown in Fig. 17-75 (c). When the aerial is pointing to the right of the target the right pulse is greater than the left pulse, and vice versa.



(a) (b) (c)

Fig.17-75 Presentation of aerial signals with electronic lobe switching (a) Input from aerial 1 (b) Input from aerial 2 (c) Input from aerials 1 and 2 alternately.

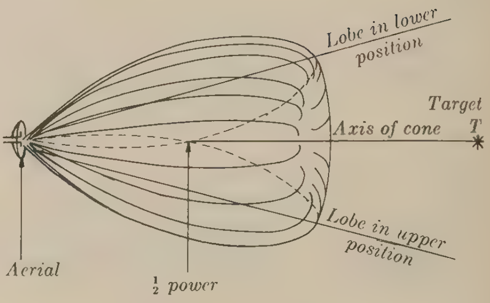


Fig.17-76 Conical scanning; aerial system directed at target.

When the aerial is pointing directly at the target, the two echo pulses have the same height.

17.7 Conical scanning. The principle of lobe switching to produce accurate indication of the bearing of a target may be extended to determine the elevation of a target as well as the azimuth. This is of particular importance in fire control designed to operate against aircraft. The directional lobe of the aerial is rotated so that its axis traces out a cone in space (Fig. 17-76). The name *conical scanning* is therefore given to this operation.

The aerial is directed in bearing and azimuth so that the axis of the cone coincides with the line from the aerial to the target. The amplitude of the echo from the target then remains constant as the lobe is rotated. If the aerial is shifted so that the axis of the cone does not coincide with the line from the aerial to the target, the amplitude of the echo pulse will vary as the lobe is rotated. If, for example, the aerial is elevated slightly, a larger echo will be received when the lobe is at the bottom of its sweep than when it is at the top. The relative phase and magnitude of the signal variation may be used to cause the aerial to be redirected to bring the axis of the cone back to the line from the aerial to the target.

The simplest method of rotating the directional lobe is to use an off-centre dipole as shown in Fig. 17-77. The dipole assembly is bent away from the axis of the reflector and rotated at a speed of 20 to 60 revolutions per minute by a scan motor. This method has the disadvantage that the dipole assembly is not balanced mechanically about the axis of rotation. Several methods of phasing the elements of the aerial to produce a lobe which is off the axis of symmetry of the aerial have been devised to overcome this difficulty. They are similar in operation to the aerials already described for lobe switching.

Somewhat different methods of producing the conical scan are used with waveguide feed. One method is shown in Fig. 17-78. It consists essentially of a rotating polystyrene-filled circular waveguide placed in the reflector end of a fixed circular waveguide. The rotating waveguide is bent so that its axis does not point directly at the centre of the reflector. Hence the directional lobe of the aerial assembly is displaced from the axis of the reflector end of the waveguide.

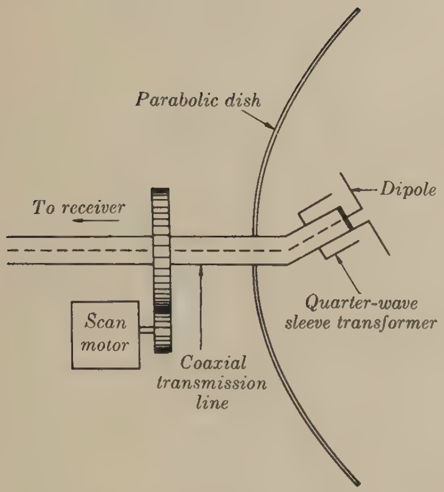


FIG.17-77 Dipole scanner.

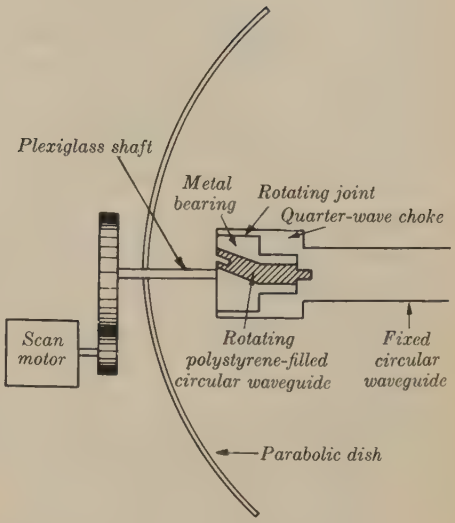


FIG.17-78 Waveguide scanner.

The polystyrene-filled guide is coupled by a plexiglass shaft to the scan motor which rotates the guide and hence the lobe. The fixed guide is matched to the rotating guide by the shape of the inner end of the polystyrene. The rotating waveguide is matched to the reflector and free space by a small hole in the outer end of the polystyrene.

The rotating joint is a metal-to-metal bearing between the main waveguide and the rotating waveguide. Loss of energy through this joint is prevented by a groove around the inner surface which acts as a quarter-wave choke.

17.8 Questions and problems.

1. Draw a block diagram of a radar set and explain briefly the purpose of each of the fundamental components.
2. Why is it desirable to employ a very-high-frequency r-f carrier?
3. What additional difficulties are presented by the use of a very high frequency for transmission?
4. What determines the minimum and maximum permissible values of pulse repetition frequency?
5. What is meant by the transmitter pulse width?
6. What determines the maximum permissible pulse width?
7. What is meant by the term duty cycle?
8. How is the duty cycle related to the peak and average values of transmitter power?
9. Distinguish between long-wave and short-wave radar frequencies.
10. What approximate frequencies and wavelengths are associated with the term microwave radar?
11. What is the purpose of the timing unit in a radar set?
12. What is the advantage of employing a separate timing unit?
13. What is meant by power-supply timing?

14. How are several radar sets prevented from interfering with each other when all are timed by the same power supply?

15. How may the transmitter be used to provide timing of the set?

16. Why is indicator timing required when calibration markers are used for ranging?

17. What is the reason for operating the timing oscillator at a frequency of 81.95 Kc/s?

18. How is the pulse repetition frequency obtained in such a set?

19. Name the type of radar transmitter shown in Fig. 17-7. How is the transmitter maintained in an *Off* condition between pulses?

20. (a) Explain how the modulator of this set turns the transmitter on.
    (b) What determines the duration of the transmitter pulse?

21. In the circuit of Fig. 17-8, what is the purpose of
    (a) the discharge line
    (b) the thyratron?

22. (a) What is the purpose of the driven stage?
    (b) Why is it called a bootstrap circuit?

23. Why is a cathode-loaded amplifier preferable to an anode-loaded circuit in a cascade pulse amplifier?

24. Explain how the modulator of Fig. 17-8 turns the transmitter on and off.

25. What are
    (a) the voltage requirements
    (b) the capacitance requirements of C_6 in Fig. 17-8?

26. In what respect is a thyratron modulator superior to a hard-valve modulator for a magnetron transmitter?

27. (a) What is the purpose of the trigger circuit of Fig. 17-11?
    (b) Explain the operation of this circuit.

28. How is the modulator of Fig. 17-11 biased off between pulses?

29. (a) What is the purpose of the discharge line in Fig. 17-11?
    (b) Explain how it becomes charged and how it becomes discharged.

30. Describe the voltages that exist across the primary winding of T_3, the pulse cable and the secondary of T_4
    (a) between transmitter pulses
    (b) during the transmitter pulse in Fig. 17-11.

31. (a) Make a drawing of a fixed-spark-gap modulator.
    (b) How is the gap made to conduct?

32. What causes the gap to stop conducting at the end of the pulse?

33. Compare the spark-gap and the thyratron modulator as to the maximum voltage and current which each supplies to the transmitter.

34. (a) What limits the effective life of the fixed-spark-gap modulator?
    (b) How is this life extended by the rotary spark gap?

35. How is the rotary-spark-gap modulator made to conduct?

36. What is meant by the term noise
    (a) in a radar receiver
    (b) on a c.r.t. sweep?

37. Name the three chief sources of noise produced in a radar receiver.

38. Why is the initial stage of the receiver the most important one so far as the generation of noise is concerned?

39. (a) What is meant by thermal agitation?
    (b) How does it develop noise voltages?

40. Why is it necessary to have a very wide bandwidth in radar receivers?

41. What disadvantage does a wide bandwidth have?

42. What is the minimum receiver bandwidth required to amplify pulses of .5 microsecond and 1 microsecond duration respectively?

43. Name the stages and state the valve types that may be used in a receiver for long-wave radar and for microwave radar respectively.

44. Why are r-f amplifiers not used in microwave radar sets?

45. What type of valve is most suitable for r-f amplification in long-wave radar receivers?

46. Compare the local-oscillator valves in long- and short-wave radar receivers.

47. Compare the first detectors in long- and short-wave radar receivers.

48. Make a diagram of a mixer unit containing a crystal first detector.

49. Trace the path of r-f, i-f and d-c components of the rectified current in the mixer.

50. How is a wide bandwidth obtained in the i-f amplifier stages?

51. What type of valve and what type of coupling are used in video voltage amplifiers?

52. What is the purpose of a cathode-follower video amplifier?

53. Explain what is meant by blocking of the receiver.

54. What is meant by the recovery time of the receiver?

55. Explain why the receiver is insensitive to echo pulses during the recovery time.

56. Draw a diagram of a circuit used to reduce the recovery time and explain its operation.

57. Explain how gating the receiver may be used to prevent the transmitter pulse from blocking the receiver.

58. What is wave clutter?

59. (a) What is a swept-gain circuit?
    (b) From what part of the circuit does it receive its input?
    (c) To what part of the circuit is the output applied?

60. (a) Draw an anti-wave-clutter circuit.
    (b) Explain how the maximum amplitude of the output voltage is controlled.
    (c) How is the duration of the output voltage controlled?

61. (a) What is meant by a logarithmic receiver?
    (b) How is the logarithmic response achieved?

62. (a) Draw a diagram of a logarithmic amplifier.
    (b) Explain its operation when receiving small signals and when receiving large signals.

63. For what application is automatic gain control of the receiver required?

64. Make a block diagram of an *a.g.c.* circuit and explain briefly the function of each section.

65. (a) What is the purpose of the range notch?
    (b) What is the purpose of the target selector?
    (c) How are these two related?

66. What is the purpose of the detector in an *a.g.c.* circuit?

67. How is the amplitude of the tracked pulse made to increase to offset fading?

68. What is the purpose of a T-R switch in radar systems?

69. Make a diagram of a spark-gap T-R switch and describe its operation during transmission and during reception of signals.

70. What is the purpose of an anti-T-R switch?

71. Make a diagram of an aerial system with an anti-T-R switch and explain its operation during transmission and reception of pulses.

72. What characteristics of the transmitter make it necessary to use a transformer between the T-R switch and the aerial system?

73. (a) Draw a diagram of a coaxial half-wave resonant transformer containing a T-R switch.
    (b) Explain its operation during transmission and reception of pulses.

74. What is the purpose of the keep-alive electrode in a T-R switch?

75. (a) How is the cavity type of T-R switch tuned?
    (b) What is its equivalent circuit?

76. How is energy coupled from a coaxial transmission line to the T-R switch cavity?

77. How is a T-R cavity coupled to a waveguide?

78. (a) Draw a T-R cavity coupled to a waveguide by means of a branch waveguide.
    (b) Explain how it prevents high-power pulses from entering the branch and permits low-power pulses to pass.

79. What is the purpose of an echo box?

80. What is meant by ringing time?

81. Explain how the echo box is used
    (a) with type *A* presentation,
    (b) with *PPI* presentation.

82. Why must the echo box be thoroughly de-tuned when the set is being operated to detect targets?

83. Show by means of a polar diagram of a directional receiving aerial
    (a) the region of maximum sensitivity
    (b) the region of maximum sensitivity to changes of bearing.

84. How is lobe switching used to obtain greater sensitivity to changes of bearing than that possible from a single-lobe aerial system?

85. Make a block diagram of a lobe-switching aerial system using a single receiver and indicator.

86. What is the appearance of the c.r.t. pattern
    (a) when the aerial is pointing to the right of the target
    (b) when the aerial is pointing directly at the target?

87. Explain how off-centre feed from aerial to receiver has the effect of producing an off-centre lobe of reception.

88. What would be the effect of reducing slightly the effective length of the phasing section of Fig. 17-72?

89. What is the chief advantage of electronic switching of the aerial circuit?

90. Why must the electronic switch output be applied to the c.r.t. circuit as well as to the gated amplifiers?

91. Sketch the appearance of the c.r.t. pattern
    - (a) when the aerial is pointing to the right of the target
    - (b) when the aerial is pointing directly at the target.

92. What is the purpose of using conical scanning?

93. What is the effect on the receiver input
    - (a) when the conical scanner is pointing directly at the target
    - (b) when the axis of the cone is slightly above the aerial-target line?

94. How is conical scanning accomplished in the case of waveguide feed to a paraboloidal reflector?

CHAPTER XVIII

LINE COMMUNICATIONS

The bulk of commercial communications and a high percentage of Service communications ashore are accomplished by means of *closed wire circuits* or *land lines.* While it might appear that the subject of line communications falls outside the scope of a textbook on radio, it must be appreciated that both types of communication are frequently employed in forwarding a particular message. It is therefore essential that Service personnel have an understanding of all methods of communication in current use, including telephone, telegraph and teletype by both land line and wireless links. In addition, many of the techniques used in line communications are also employed in ships. The present chapter deals with the following aspects of line communications:

(i) voice-frequency telephone
(ii) d-c telegraph
(iii) carrier telephone
(iv) carrier telegraph.

18.1 Line communications versus radio communications. Line communications are, in general, more economical and more reliable than radio communications between permanent installations ashore. Radio equipment, especially transmitters, consume more electrical power than corresponding wire facilities. Also, the total number of radio channels which can be utilized is limited by frequency assignment and interference considerations, whereas the total number of wire circuits is limited only by the amount of material and personnel available for the installation and maintenance of the lines. Because of ionospheric considerations and other vagaries of propagation, radio communication ranges are less predictable than those of wire lines. Radio communications are readily subject to interception, direction finding and jamming by hostile stations and the volume and geographical distribution of message traffic may be readily ascertained by the enemy. On the other hand, wire communication is subject to interruption and, to a small degree, interception at any point along the wire route as a result of physical damage from bombing, shell fire and sabotage; whereas in radio, only the aerials and terminal equipment are vulnerable to physical damage.

18.2 Telephony versus telegraphy. The greatest advantage of the telephone is that it affords immediate personal contact between individuals. Its chief disadvantages are the lack of permanent records of conversations, the tendency to consume too much time in talk and the possibility of non-compliance with security regulations. Telegraphy is best suited to one-way messages and orders which require no explanation. A telephone can transmit information faster than an ordinary single telegraph channel since a fast talker can read 150 to 200 words per minute of ordinary material so that it can be readily understood over a good telephone circuit; the normal speed for sending over a telegraph circuit from a previously prepared tape is 60 words per minute while the speed for manual operation varies from 10 to 40 words per minute depending on the skill of the operators. The simplicity of telegraph signals as compared with the more complex speech currents makes possible the transmission of 12 telegraph

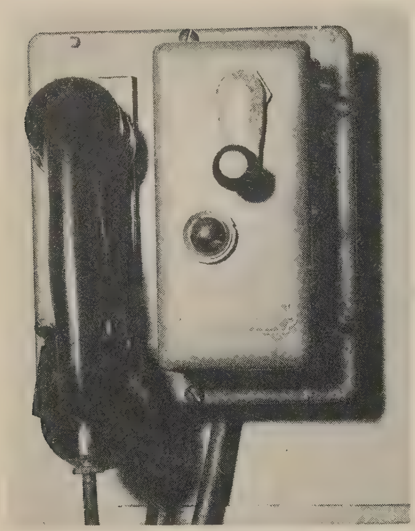


FIG.18-1 Single unit of sound-powered telephone link.

messages over a frequency band no wider than that required for one telephone conversation.

18.3 Telephone systems. A complete telephone system includes speech transmission channels, telephone instruments and switching and signalling equipment. A telephone transmission channel is a path which carries a single conversation between two telephones. For satisfactory transmission, the received speech on this channel must be loud enough to be understood and, in addition, certain unwanted effects such as noise, crosstalk and the tendency of repeaters to sing must be controlled.

Telephone systems may be roughly divided into two types, *sound-powered* and *battery-powered*. Sound-powered telephones are operated by the voice without any external power supply; they have a much lower transmitting efficiency and can therefore be used only on relatively short circuits and when the transmission loss between telephones is small. Battery-powered telephones may use batteries located at the telephone (local battery system) or may obtain battery supply over the loop from the telephone central office (common battery system).

A. Sound-Powered System

Sound-powered telephones are widely used in shipboard installations where the distance between units is small and consequently the line loss is low. Two units of the type shown in Fig. 18-1 make up a system for point-to-point communication. Sound-powered telephones may also be used on switchboard connections such as that shown in Fig. 18-2. This unit contains two operator's handsets, signalling lamps, switches for connecting together any pair of sixteen outstations and ringers for calling these stations. A switchboard installation for use in larger ships and shore establishments is shown in Fig. 18-3. Power to operate the signalling lamps and relays contained in the switchboard unit is supplied by a suitable d-c source while the power required to operate the handsets is provided by the voice.

The handset of the sound-powered unit is similar in appearance to the battery-operated unit and contains a microphone and a receiver. No current is supplied to the microphone of the sound-powered unit however. The microphone contains a permanent magnet, coil and diaphragm and is similar in construction

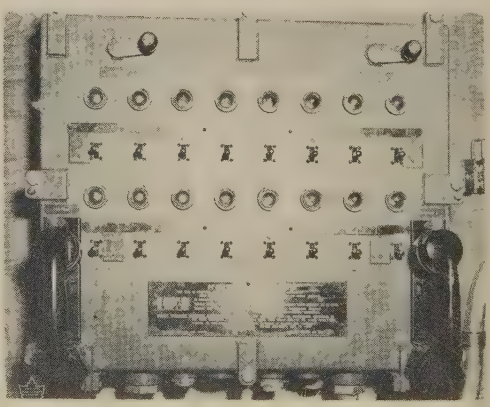


FIG.18-2 Switchboard unit of sound-powered telephone system.

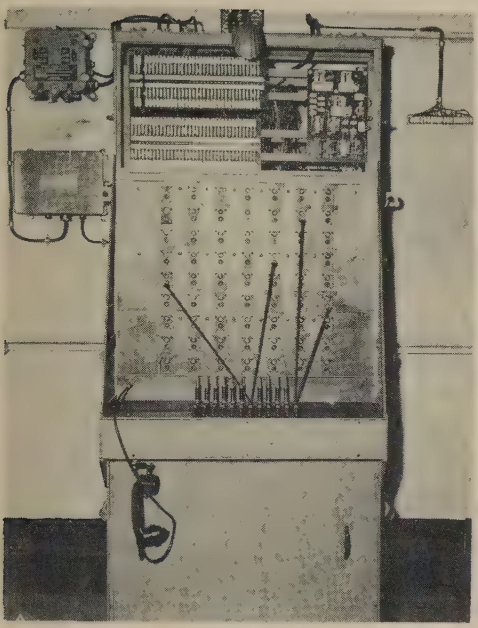


FIG. 18-3 Telephone switchboard unit for large ships and shore installations.

and operation to a headphone. Sound waves, entering the microphone, set the diaphragm in vibration and set up induced currents in the coil. These currents are made to flow through a closed wire loop to the receiver of the remote unit. The receiver is a resonant magnetic unit in which the diaphragm is undamped and clamped at the edge. Mechanical coupling is used between the diaphragm and the armature to prevent excessive loss in efficiency. The voice current from the microphone sets the armature and diaphragm in vibration and reproduces the sound waves entering the microphone.

The efficiency of a sound-powered link is about 25 *db* lower than that of a similar carbon microphone system. This limits the sound-powered equipment to point-to-point applications where the line loss does not exceed about 15 *db*. Because of the resonant characteristics of the microphone and receiver units, the frequency response of the link falls off sharply for frequencies above 2000 c/s. The signalling equipment used with sound-powered units is generally of the hand-operated type as indicated by the ringers of Fig. 18-1 and Fig. 18-2.

B. Common-Battery System

Most telephone service in urban areas, private branch exchanges (*PBX*), and powered shipboard telephone installations are of the common-battery type. A simplified diagram of a common-battery telephone system is shown in Fig. 18-4.

The network consists of two outstations operated by parties *A* and *B* and a switchboard attended by a switchboard operator. Each outstation is connected by a two-wire line to the switchboard. Associated with each outstation are two signalling lamps on the switchboard panel. The *line lamp* indicates to the switchboard operator that the outstation operator wishes to contact him. The *supervisory lamp* indicates to the switchboard operator that the outstation operator has terminated the conversation and that the line may be cleared. The following summary outlines the sequence of events which take place in initiating and terminating a call from party *A* to party *B*.

(i) *A* removes receiver from hook causing line lamp to signal switchboard operator.
(ii) Operator inquires from *A* the party to be called.
(iii) Operator connects line to party *B* causing supervisory lamp to light.
(iv) Operator rings *B*.
(v) *B* answers, causing supervisory light to be extinguished.
(vi) When conversation is concluded, *A* and *B* hang up receivers, causing both supervisory lamps to light.
(vii) Operator removes connecting link from line *A* to line *B*. This extinguishes supervisory lamps.

A detailed explanation of the operation of the circuit is given in the follow-

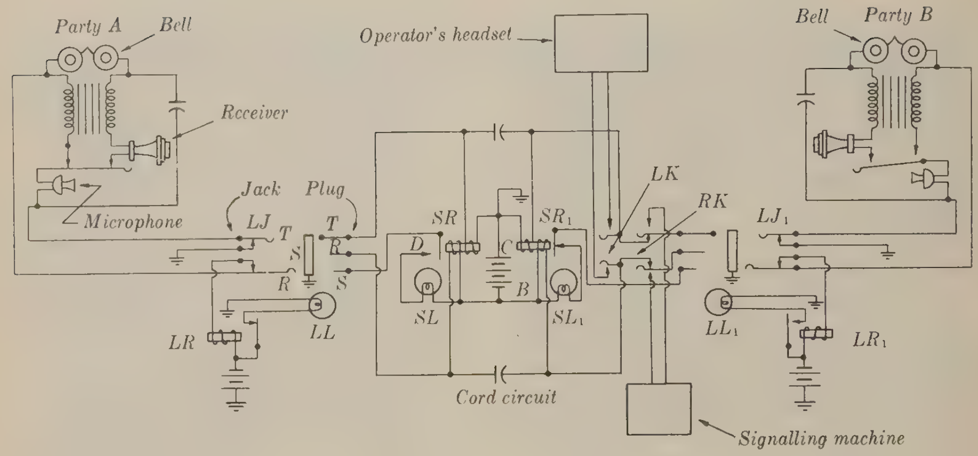


FIG. 18-4 Simplified circuit of a non-multiple common-battery system.

ing paragraphs. If party *A* wishes to call party *B*, he removes the receiver from the hook. The switch-hook contacts close as shown in the diagram of Fig. 18-4 allowing direct current to flow through the line relay *LR*. When energized, this relay pulls up its armature, closing the path through line lamp *LL* and thus signalling the operator.

The operator answers this signal by inserting the answering plug of a cord circuit into the line jack *LJ* of the calling party as indicated by the lighted line lamp. The insertion of this plug does three things. First, the contacts are spread apart, de-energizing the line relay *LR*, allowing its armature to fall back under the force of gravity and thus extinguishing the line lamp. Second, the tip contact *T* and the ring contact *R* of the plug make contact with *T* and *R* of the line jack. The making of these two contacts permits direct current from the common battery *CB* to flow out to the calling telephone *A*. The path of this electron flow is from the negative battery terminal through one half of the inductive supervisory relay *SR* to the ring of the plug; then through the telephone set, back through the tip of the plug, through the other half of the supervisory relay to the positive (grounded) battery terminal. The current through the two halves of the supervisory relay *SR* causes the armature to open contact *C*. Third, the sleeve contact *S* on the plug touches the grounded contact *S* on the jack. The supervisory lamp *SL* would now light if the supervisory relay had not previously opened. The circuit of the supervisory lamp is, however, now in a condition to light as soon as this relay contact makes, as it does in a later operation.

The operator now throws the listening key *LK* to the listening position; this closes the contacts at the key and connects the operator's headset across the line. The desired number is obtained from the calling party and, if the called line is not busy, the operator inserts the plug of the calling cord corresponding to the answering cord previously used into the line jack of the called party.

When the plug is inserted, the contacts of line jack LJ_1 are opened. This prevents the line lamp LL_1 from lighting when the receiver of the called party *B* is removed from the hook. The insertion of the plug also lights the supervisory lamp SL_1 through the grounded battery and ground on the sleeve. The receiver at station B is still on the hook and no current is flowing through SR_1 to hold open the armature contact of this relay. The operator now throws the ringing key RK to the ringing position. This connects the called line to the signalling supply and opens the contacts to the line of the calling party so that party A will not receive the unpleasant signalling sound in his receiver.

When called party B removes the receiver from the hook, the contacts at the receiver make and current from the battery flows out to the called station B through the two halves of the supervisory relay SR_1. The current actuates the armature of this relay, thus breaking the contact and extinguishing the supervisory lamp SL_1. This indicates to the operator that the party has answered and that signalling may cease. The switchboard lamps are not lighted while the conversation is in progress. Conversation takes place through the modulation of the direct current from the battery by the sound waves at the transmitting device.

When the conversation has been completed, A and B replace their receivers on the hooks. This breaks the receiver-hook contacts at both sets, thus breaking the path of the current through SR and SR_1 respectively. The armatures of these relays drop back closing the associated contacts and permitting current to flow through each of the lamps SL and SL_1. The lighting of these lamps indicates to the operator that the conversation is completed and the connection is cleared by removing both plugs from the jacks.

C. *Multiple Switchboards*

In a small exchange, where a single operator can handle all incoming lines, it is possible to connect any two stations together when each station's line terminates in a single jack only. However, when there are more than a few hundred stations, the lines of which terminate in the same switchboard, it is impossible for one operator to handle all the lines. In order to apportion the work and to make it possible to mount the switchboard apparatus in such a way that it will permit interconnection of a large number of lines, the multiple switchboard was introduced. In a multiple switchboard, the answering jacks and signals are divided among several operators; each operator handles, on the average, about two hundred lines and is responsible for answering any signals from these stations. In addition to these *answering jacks*, there may be as many as 3,300 *calling jacks* in the position in front of each operator. Calling jacks do not have signal lamps mounted with them as in the non-multiple board, thus allowing a large number of them to be mounted in a relatively small space. The calling jacks are multipled, that is, connected in parallel with similarly located jacks at every third position down the board. In this manner, about 10,000 calling jacks are within reach of any operator, either in the position directly in front or in the adjacent position to the left or right. The operator is warned against plugging into a busy line by means of a click which is heard in the operator's headset when the answering plug is inserted into a calling jack already in use. Fig. 18-5 illustrates a typical arrangement of jacks in a common-battery multiple switchboard.

Since about 10,000 multiple jacks are all one operator can conveniently reach, a system with more than this number of telephones must have more than one central exchange. When a party connected to one exchange wishes to talk with one connected to another exchange, the call must be connected by means of inter-exchange *trunks* for completion. Each operator must accordingly be sup-

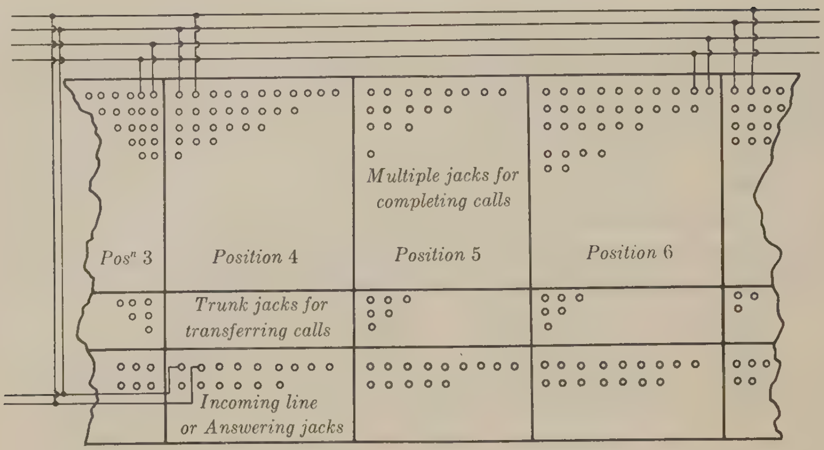

Multiple jacks for completing calls
Posn 3
Position 4
Position 5
Position 6
Trunk jacks for transferring calls
Incoming line or Answering jacks

FIG.18-5 Arrangement of jacks in a common-battery multiple switchboard.

plied with trunk jacks to each of the other central exchanges. These trunk jacks are also shown in Fig. 18-5.

D. Signalling

To provide a means for the operators at the two ends of a circuit to signal each other and to signal parties called, signalling equipment or *ringers* are required for each circuit. Signalling on line circuits is accomplished by means of direct current, 20 c/s alternating current or voice-frequency current. Direct current is used for signalling on loops from telephones to common-battery switchboards and on automatic signalling trunks. Twenty c/s current is used to ring bells at telephones and to ring operators over magneto loops. Carrier telephone channels and certain types of repeaters do not transmit 20 c/s current; on these circuits, voice-frequency signalling must be used. A voice-frequency ringer converts the 20 cycles per second a-c received locally to a frequency (usually 135, 500 or 1000 c/s) which will pass freely over the telephone channel. This frequency is ordinarily interrupted at regular intervals to provide a 20 c/s signal which can be distinguished from voice currents in the remote receiver.

In present-day telephone practice, automatic ringing is used. That is, a ringing signal is sent automatically at regular intervals until the called party answers. The ringing tone, very much reduced in volume, automatically advises the calling party when ringing is in process. An audible busy signal tone is automatically placed on the calling line when the called line is busy. These changes have resulted in a considerable improvement in the service.

E. Automatic Dial Systems

Automatic dial systems, often called *machine switching systems*, are a means of connecting desired lines together at the central office by purely mechanical methods. The first such system, the *step-by-step dial system*, was developed by Strowger in 1887 and, in a modified form, is still in wide use. In some large telephone plants, the step-by-step system has been replaced by the *crossbar system* which was first introduced on a practical basis just prior to the Second World War. Dial systems are quicker and more accurate than manual service and have proven more adaptable to improvements and developments in the telephone industry.

1. STEP-BY-STEP SYSTEM

The step-by-step system of dial operation is best adapted to telephone systems or exchange areas having a relatively small number of central offices. The switching mechanisms are operated by the telephone user by means of the dial on his telephone instrument. When the dial is turned, the dial contacts open the receiver connection and short-circuit the transmitter. As the dial is released, the current through the line is interrupted at regular intervals sending current impulses to the central office to operate the switching mechanisms. The speed of rotation of the dial as it returns to its normal position is accurately controlled by a governor. This mechanism provides a series of well-timed electrical impulses to the line.

A schematic diagram for a small private automatic exchange (*PAX*) capable of providing step-by-step dial operation for 100 lines is shown in Fig. 18-6. It consists of two types of switch, a *line switch* or *line finder* and a *connector switch*. The line finder is used to reduce the number of lines required to provide communication between the 100 operators using the network. If the line finder were not used, it would be necessary to have one connecting line for each of the 100 operators. This would not be practical since only 50 simultaneous calls are possible while under normal circumstances a much smaller number would be in actual progress at any one time. In practical installations, therefore, each line terminates in a line switch and for each 50 line switches, nine trunks leading to nine connector switches are provided. The connector switch is a two-motion switch bank assembly of the Strowger design. It contains 100 contacts mounted in 10 horizontal rows of 10 contacts each. Each contact is insulated from the others and is connected to a terminal block at the back of the switch. Electrical connection may be made to any of the 100 contacts by means of a wiper attached to a vertical switch shaft. The shaft may be raised by means of an electromagnet to bring the wiper to the vertical level of any of the ten contact banks. The shaft may then be rotated by another electromagnet to cause the wiper to make an electrical connection with any of the ten contacts in the horizontal bank selected.

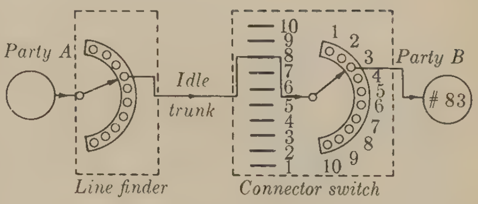


Fig. 18-6 Schematic diagram of a 100-line step-by-step dial telephone system.

The operation of the circuit of Fig. 18-6 is best understood by means of a particular example. Suppose that calling party *A* wishes to call party *B*, number 83. When *A* removes the handset of his unit, the line finder switch automatically connects his line to a trunk leading to an idle connector switch. This occurs in a fraction of a second and without the knowledge of the calling party. When this connection is completed, the calling party receives dial tone. He then dials the first digit, 8. This operation sends a series of 8 impulses to the line which cause the switch shaft in the connector switch to step up to the eighth level. Operator *A* then dials the second digit 3 which sends 3 impulses to the line. These impulses operate the rotary mechanism which causes the wiper to move to horizontal position 3 on vertical bank number 8. This completes the connection between the two telephones and, if the called line is not busy, the telephone of party *B* is rung. The transmission circuit is completed when the called party answers. When both parties replace their handsets, the switches are automatically released and the lines are cleared for another call.

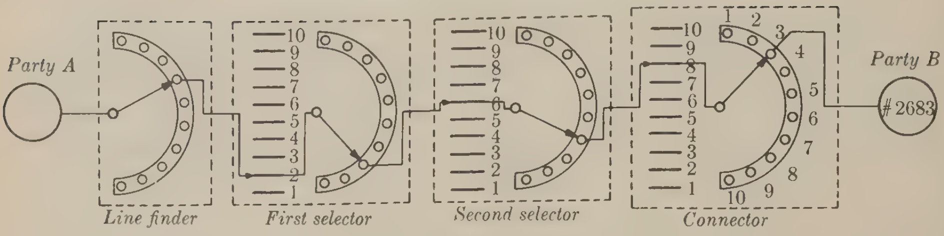


FIG. 18-7 Schematic diagram of a 10,000-line step-by-step dial telephone system.

A schematic diagram of a 10,000-line step-by-step dial system is shown in Fig. 18-7. It is similar to that of the 100-line system of Fig. 18-6 but has two additional switches called *selectors* connected in series between the line finder and the connector. The selector is similar in construction to the connector but its operation is a combination of the connector and line finder operations. Suppose party *A* wishes to call party *B* at number 2683. When *A* removes his handset, the line finder switch finds a line to an idle first selector switch. *A* then hears dial tone. When he dials 2, the first selector switch shaft elevates the wiper mechanism to the second level. This connects party *A* to the 2000 group of lines. The shaft then rotates to find a line to an idle second selector. In this respect the selector is similar to the line finder. When operator *A* dials 6, the second selector operates, elevating the wiper to the sixth level. This connects *A* to the 2600 group of lines. The shaft then rotates to find a line to an idle connector switch. The operation of the connector circuit caused by party *A* dialling the remaining two digits 8 and 3, is precisely the same as that described for the 100-line system. It is evident that the number of lines that can be connected by the step-by-step equipment may be further increased by increasing the number of selector switches in the chain. This service is provided in large cities where the entire subscriber population is divided into regions or exchanges. Dialling the first number in a 5-number system, or the first two letters of the exchange name as the case may be, connects the subscriber to the desired exchange. Thereafter the sequence of operations is as described for the 10,000-line system.

In addition to the functions described above, the selector and connector switches may be required to carry out some or all of the following operations:

(i) to hold all preceding switches in the switch train operative until the connection is released

(ii) to return *busy tone* to the calling party if all selectors are busy or if the called party line is busy

(iii) to ring the called party and to return *ring-back tone* to the calling party thereby indicating that the called line is being rung

(iv) to remove ringing current from the line when the called party answers

(v) to supply transmission current to both parties

(vi) to release all switches and clear the lines when the call has been completed and the parties replace their handsets.

2. CROSSBAR SYSTEM.

In the step-by-step system, wiping contacts are used and constant wear results. A wiper may have to sweep over as many as nine contacts to reach the required one. Wiping contacts are eliminated in the crossbar system by mounting a pair of contacts at each position in a bank of connections. To make a connection between contacts it is only necessary for a simple mechanical move-

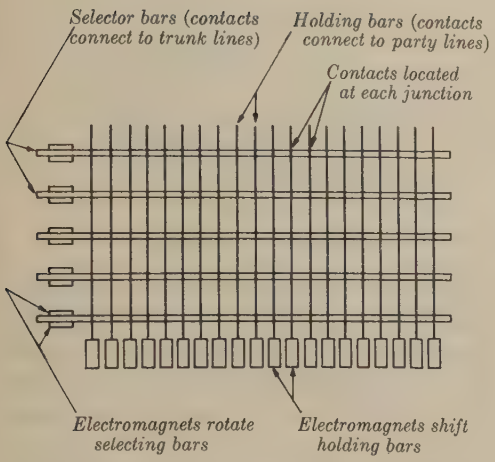

Selector bars (contacts connect to trunk lines)
Holding bars (contacts connect to party lines)
Contacts located at each junction
Electromagnets rotate selecting bars
Electromagnets shift holding bars

FIG.18-8 Schematic diagram of crossbar switch unit.

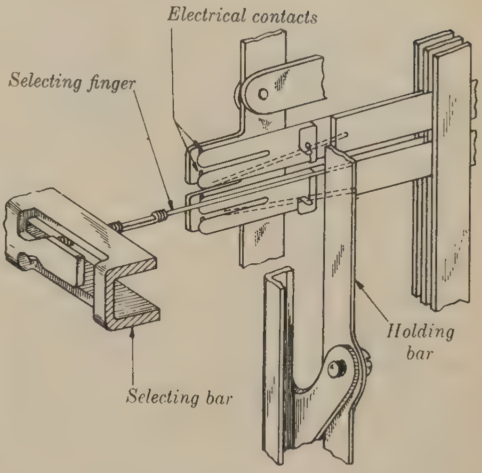


FIG.18-9 Crossbar switch.

ment to press the paired contacts together.

A single 200-point crossbar switch unit is shown schematically in Fig. 18-8. It consists of 5 horizontal *selecting bars* and 20 vertical *holding bars* mounted in parallel planes in an upright frame. The 5 selecting bars may be rotated or twisted in either of two directions by means of electromagnets, two magnets being associated with each bar. Two pairs of contacts are mounted at each of the 100 junctions of the horizontal and vertical bars. These contacts may be pressed together to make an electrical connection at any intersection by rotating the correct selecting bar through a small arc and then actuating the proper holding bar. The contacts associated with the 20 holding bars are connected to 20 party lines while the contacts associated with the 5 selecting bars are connecting to 10 trunk lines. Thus, any of the 20 party lines may be connected to any of the 10 trunk lines by energizing one of the holding bar magnets and one of the selecting bar magnets.

The mechanical arrangement of the selecting bars, holding bars and contacts at each junction is shown in Fig. 18-9. A *selecting finger*, made of wire, projects in a horizontal direction at right angles to the selecting bar. One end of the finger is mechanically coupled to the selecting bar while the other end normally lies between the two pairs of electrical contacts. When the selecting bar is rotated through a small arc by one of its electromagnets, the selecting finger is moved up or down to lie across one U-shaped metal connecting strip which is normally held away from the two contact points by spring action. If now the electromagnet which controls the holding bar is energized, the holding bar presses the connecting strip against the contacts thus connecting the contacts. At those junctions where the finger had not been previously actuated by the selecting bar, the holding bar merely forces the finger into the space between the strips and between the sets of contacts but does not press against the strips.

The holding bar is kept in its operating position for the duration of the call, but the selecting bar returns to normal immediately after the holding bar has operated. This permits all the fingers to return to their horizontal positions except the one actuated by the holding bar.

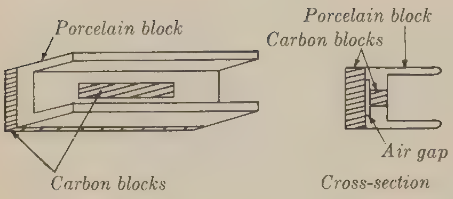


FIG.18-10 Open-space cutout.

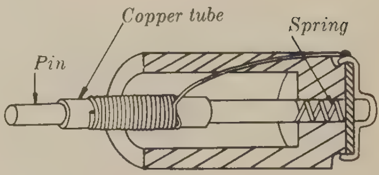


FIG.18-11 Heat coil.

The selecting bar is then left free for another selection with a different holding bar. Provision for connecting together more party lines and trunks is made by interconnecting a number of crossbar units.

The crossbar dial system provides a greater flexibility in trunking arrangements than the step-by-step dial system. This is due not only to the simple crossbar switching system described above, but also to the *marker system of control* which is used in establishing connections throughout the crossbar office. This enables the circuit to make two or more attempts to establish a call over alternate switches and trunks when normally used paths are all busy. It is also arranged to detect short-circuited, crossed, grounded and open-circuited conditions at all vital points. Before the components are released from a connection, the circuit is checked to ensure that the connection has been properly established. When faults are detected an alarm is sounded and the location and nature of the trouble are indicated. The operating time of the markers is considerably less than one second so that only three or four markers of each type are required in an average office.

F. Protection

To protect the switchboards and associated equipment from damage which might otherwise result from lightning or from accidental contacts between communications circuits and high-voltage power lines, protectors are placed in lines coming in to telephone centrals from the outside. Protective devices are of three principal types: *open-space cutouts*, *fuses* and *heat coils*. Open-space cutouts and heat coils operate to connect the protected wire to ground while fuses open the wires in which they are inserted. Each of these protective devices is designed so that, for the particular situation in which it is used, it will be sufficiently sensitive to operate before the plant which it is protecting is damaged but not so sensitive as to cause an unnecessary number of service interruptions. The standard form of open-space cutout is shown in Fig. 18-10. It consists of two carbon blocks having a separation of a few thousandths of an inch, one of which is connected to ground and the other to the wire to be protected. One of the carbon blocks is much smaller than the other and is mounted in the centre of a porcelain block. When the voltage on the wire becomes dangerously high, the wire is grounded by arcing across the small air gap between the carbon blocks. If considerable current flows, enough carbon will be drawn into the gap by the arc to cause permanent grounding. When a telephone conductor is grounded by the action of an open-space cutout, current will continue to flow through the conductor to ground as long as the exposure continues. This current may be large enough to damage the conductor or the protective device. It is thus necessary to insert on the line side of the cutout a device that will open when the current is too large. Fuses are used for this purpose.

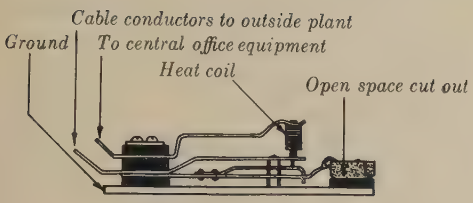

Cable conductors to outside plant
Ground
To central office equipment
Heat coil
Open space cut out

FIG. 18-12 Cutout and heat coil assembly.

Finally, it is frequently necessary to protect telephone apparatus against external effects in which the voltage is not high enough to operate the cutout or the current high enough to operate the fuse. Such currents, which will damage equipment if allowed to flow over a long period of time, are called *sneak currents* and are guarded against by the use of heat coils. A heat coil (Fig. 18-11) consists of a small coil of wire wound around a copper tube. It is connected in series with the wire to be protected. A metal pin is inserted within the copper tube and held in place by a piece of solder with a low melting point. The pin is connected to the line side of the coil. If sufficient current flows through the coil to melt the solder, the pin will move under the pressure of its mounting spring and thus connect the line to ground. A method of mounting heat coils and open-space cutouts on the protector frames at central offices is illustrated in Fig. 18-12.

G. Plugs, Jacks and Cords

Plugs, jacks and cords are widely used in telephone and telegraph equipment as a means of performing switching operations rapidly and with a maximum of flexibility. The mechanical and electrical features of a two-conductor plug are shown in Fig. 18-13. The plug consists of a central brass rod ending in a round tip. A brass sleeve surrounds the rod for most of its length and is insulated from the rod and tip by a bakelite bushing. Connection is made to the rod and sleeve by means of brass screws. A bakelite handle surrounds the connecting unit and facilitates handling the plug.

A three-conductor plug and jack assembly is shown in Fig. 18-14. The jack consists of a metal frame with a brass sleeve fastened to the front face. Spring contacts and auxiliary contacts are insulated from each other and mounted on the frame. When the plug is inserted into the jack, electrical connection is made from the three conductors to the sleeve, ring and tip contacts of the jack. Auxiliary contacts may be mounted in such a way as to open or close other circuits when the plug is inserted. The construction of a common type of three-conductor switchboard cord is shown in Fig. 18-15. Cords, plugs and jacks of the type described above are used in manual switchboard installations of the type shown in Fig. 18-4.

H. Switchboard Keys

A telephone key is a type of switch specially adapted for use in telephone or telegraph circuits. Such a key may perform the same circuit functions as a double-pole double-throw knife switch as

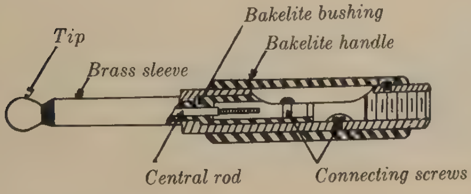


FIG. 18-13 Cut-away section of a two-conductor telephone plug.

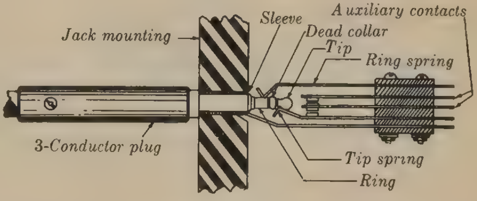


FIG. 18-14 Three-conductor telephone plug and jack assembly.

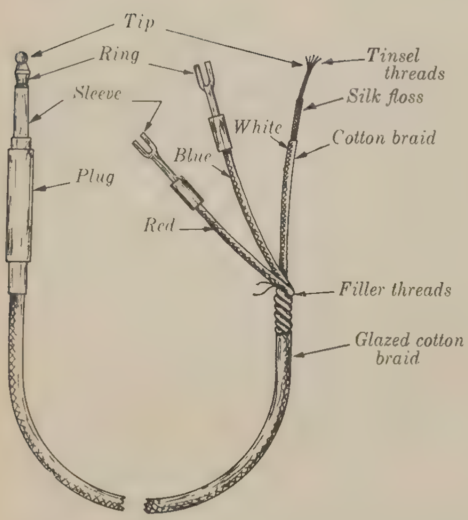


Fig. 18-15 Construction of three-conductor switchboard cord.

shown in Fig. 18-16. The key-type construction permits contacts to be made or broken by means of carefully adjusted German silver springs. Contact metal is welded to the springs to prevent excessive resistance from being introduced into the circuit. The ringing keys referred to in Fig. 18-4 are usually of this type.

Fig. 18-17 illustrates a key assembly widely used in the operation of switchboard circuits. It is constructed on the unit principle which permits one or more key spring units to be mounted on a standard metal base. Additional contacts may be included in the assembly for opening or closing auxiliary circuits along with the main circuits.

18.4 Telegraph circuits. Communication by telegraph depends upon the transmission of electrical signals arranged according to a code. These signals are formed by interrupting or reversing the direction of a continuous current in accordance with the signal code used. When the key is closed at the transmitting station current is sent out to the line. The circuit is then said to be producing a *mark* and this condition is called *marking*. When the transmitting key is released, the current in the line is cut off (or reversed in certain systems)and the circuit is said to be producing a *space*. This condition is called *spacing*.

Wire telegraph systems may be divided into two categories, carrier circuits and d-c circuits. Carrier telegraph facilities operate in the voice-frequency range and are generally used for long-distance communication on land lines. D-c circuits are used for short-distance and branch-line circuits.

A. Repeaters

The limiting length of a land line is reached

(i) when the signal becomes so weak as to be incapable of actuating the receiving apparatus

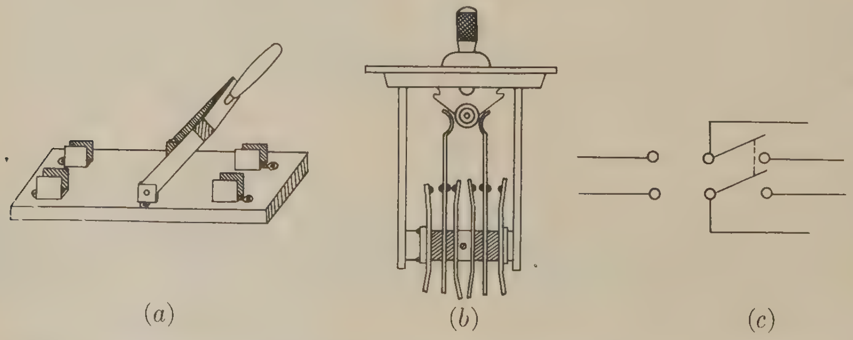


Fig. 18-16 Double-pole double-throw switches (a) Knife switch (b) Telephone type (c) Circuit symbol.

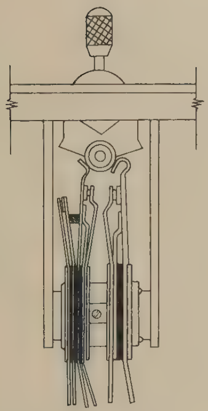


FIG. 18-17 Universal key assembly.

(ii) when the waveshape is modified so that the time distortion of the signal impulses is excessive

(iii) when the received signal strength is too low to over-ride the interference.

Usually the circuit can be extended by inserting a *repeater* before the limiting length is reached. In d-c telegraphy, such a repeater involves the reception of signals by means of a relay and automatic re-transmission using a local source of energy. In carrier telegraphy the repeater either converts the signals to direct current before re-transmission or, in the case of a valve repeater, amplifies the voice-frequency signals without conversion.

B. Single and Duplex Operation

Telegraph operation which is limited to one direction over a circuit at one time is known as *single* or *half-duplex* operation. When such a system is used, the receiver may not begin to transmit signals until the sender has terminated his transmission. With this type of operation, however, a break feature is usually provided to enable the receiver to stop the sender. A circuit in which independent transmission paths are provided for the two directions of transmission is known as a *duplex* or *full-duplex* circuit. A duplex system makes it possible to transmit signals in both directions over the loop at the same time without interference.

C. Neutral and Polar Operation

A d-c *neutral* circuit is one in which current flows through the line from the transmitting station to the receiving station during the mark condition and no current flows during the space condition. An elementary d-c neutral telegraph circuit is shown in Fig. 18-18. The line current is provided by a battery at one point in the line. Marking and spacing are effected by a hand-operated key at each end of the line for single transmission. The line may consist of a single wire with ground return, as shown, or of a pair of wires.

A d-c *polar* circuit is one in which the current flows in one direction through the line during the mark condition and

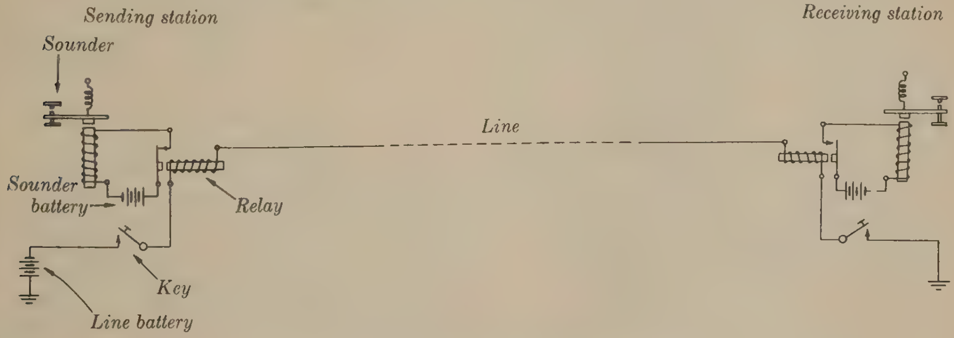


FIG. 18-18 Elementary neutral telegraph circuit.

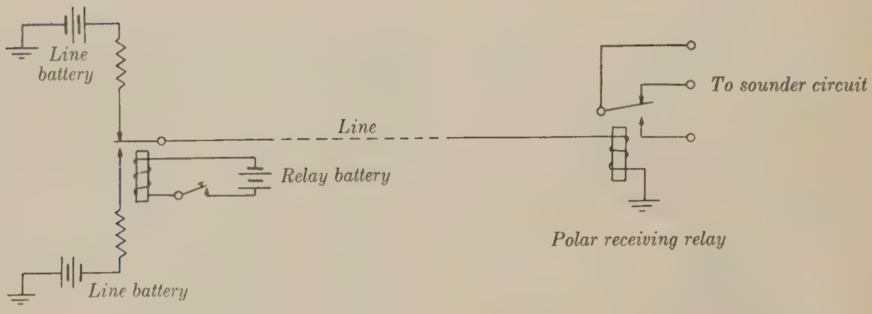


FIG.18-19 One-way polar telegraph circuit.

reverses during the space condition. Approximately equal currents flow in the two directions. This system requires two line batteries as shown in Fig. 18-19. A polar receiving relay is required at the receiving end of the line. The armature of this relay must move to the mark position for current flowing through it in one direction and to the space position when the current reverses.

A double-current system, such as the polar circuit just described, is less affected by changes in insulation resistance than a single-current or neutral system. The leakage currents passing through a neutral relay tend to keep it energized at all times, even when sending a space. Thus, the relays must be adjusted to operate on changes in the magnitude of the current. As the line leakage changes with changes in the weather, the marginal operation of the relay must be continually adjusted. No such adjustments are required in equipment operating on a polar basis to compensate for varying weather conditions since the operation of the relay depends upon changes in direction of line current rather than changes in magnitude. For this reason, the operating range of a polar circuit is considerably greater than that of a corresponding neutral circuit provided with an adjustable receiving relay.

D. Polarential Operation

A *polarential* system is one which combines some of the features of both polar and neutral operation. It permits true polar operation from the central office to an outlying point and a modified polar operation in the reverse direction. Thus the advantages of polar transmission are secured and at the same time the equipment required in the outlying office may be simplified. The essential components of a polarential telegraph circuit are shown in Fig. 18-20.

During transmission from station *A*

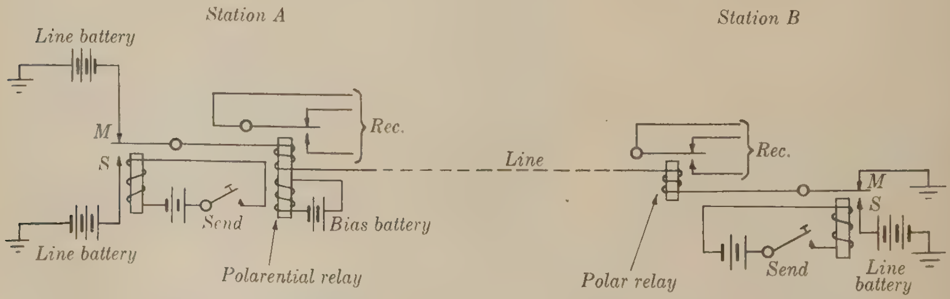


FIG.18-20 Polarential telegraph circuit.

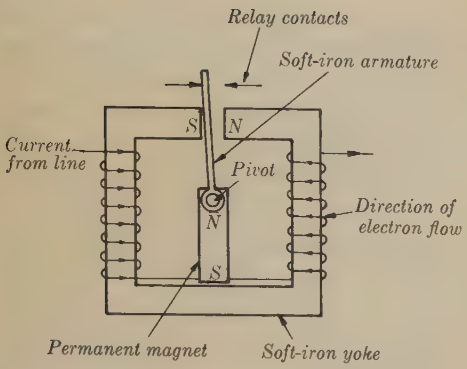


Fig. 18-21 Operation of polar relay.

(central office) to station *B* (outlying office) the line is always grounded at station *B* and normal polar operation is employed. The two line batteries at *A* alternately provide equal currents in opposite directions for the mark and space conditions. Transmission from *B* to *A* in the mark position is accomplished by connecting the line to ground at station *B*. The armature of the polar receiving relay at *A* is held in the mark position by current through a bias winding on this relay. The effect of this current on the receiving relay is to oppose and overcome the effect of the current supplied to the line by the line battery at *A*. When the transmitting key at *B* is put to the space position, the line batteries at *A* and *B* are connected in series. The effect of the increased line current on the receiving relay is to overcome the effect of the bias current and move the receiving armature at *A* to the space position.

E. Polar Relays

A *polar relay* is a relay which responds to changes in direction of current rather than to changes in magnitude. Fig. 18-21 shows in diagrammatic form one type of polar relay. It consists of a horseshoe-shaped yoke of soft iron on which the relay windings are located. Inside the yoke a polarizing permanent magnet is placed. A soft-iron armature completes the magnetic circuit. One end of the armature is pivoted at the upper end of the permanent magnet while the other end is free to move to either side of the air gap between the poles of the yoke. In the absence of any current through the winding, the magnetic flux of the permanent magnet is free to divide equally through the two halves of the soft iron yoke and there is equal chance that the armature will rest in the left or right position. If now, the line current flows through the winding as illustrated, the magnetic flux due to this current is added to that of the permanent magnet in the left branch of the yoke and opposes the flux in the right branch. The armature is therefore drawn to the left to bridge the air gap as shown. When the line current reverses the flux is greater in the right yoke and the armature moves to the right.

Another type of polar relay called a *differential polar relay* is shown in Fig. 18-22. While different structurally, the theory of operation of this relay is similar to that of Fig. 18-21. The permanent magnet is of the horseshoe type and is placed around the outside of the coil. The coil windings are suspended between

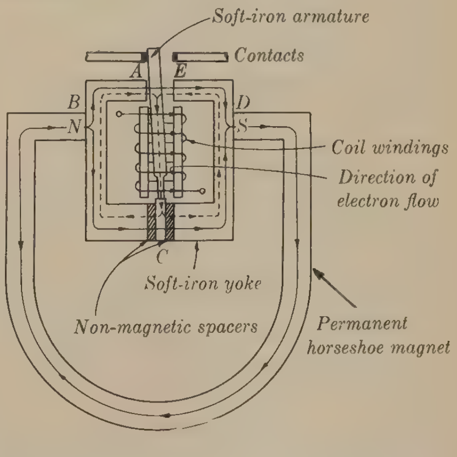


Fig. 18-22 Construction of receiving relay for polar or polarential systems.

the sides of the magnet. A soft-iron magnetic yoke surrounds the coil and completes the magnetic path of the horseshoe magnet. A soft-iron armature is suspended inside the coil and is free to move in the air gap between the pole pieces of the yoke.

In the absence of line current through the coil windings, the magnetic flux of the permanent magnet is distributed through the magnetic circuit as indicated by the solid lines and arrows. Equal forces are exerted on both sides of the armature. When electrons flow through the coil as indicated in Fig. 18-22, additional magnetic flux is set up through the soft-iron yoke as shown by the dotted lines. As a result, the total magnetic flux is increased from B to A and from C to D. The flux is decreased from B to C and from E to D. The armature is therefore drawn toward the left to complete the magnetic circuit from A to C. If the current through the coil is reversed, the armature is drawn to the right. An extra winding may be included in addition to the line winding to provide a biasing effect as explained for polarential operation.

18.5 Increasing telegraph traffic. To ensure reliable service, it is common practice to place important long-distance telegraph conductors inside a metal sheath. *Metallic telegraph systems* make use of this type of cable construction as opposed to open-wire construction with ground return. Several channels are usually provided by superimposing telegraph messages on cable pairs already carrying telephone speech currents. The cable conductors are very close together, are provided with repeaters and are loaded to extend the useful length of the line. To prevent interference with the telephone traffic, telegraph currents and voltages are maintained at a much lower level than those used on lines carrying only telegraph signals. Line currents of 4-5mA are used in metallic telegraph systems as compared to values of 30-60mA in an ordinary installation. More sensitive relays than those described in the preceding sections are required for metallic systems.

A. Simplexed Circuits

Three methods are commonly used to provide simultaneous telegraph and telephone service over the same line wires:
(i) the simplexed circuit
(ii) the composite arrangement
(iii) carrier systems.
Carrier systems are discussed in a later section. In a *simplexed circuit,* one winding of a repeater coil or transformer is connected across the line pair and a point midway between the coil terminals of this line winding is connected to the telegraph equipment as shown in Fig. 18-23.

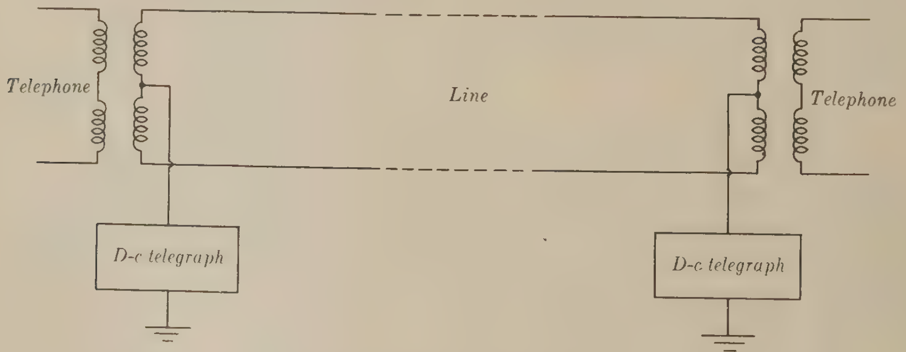


FIG. 18-23 Simplexed circuit.

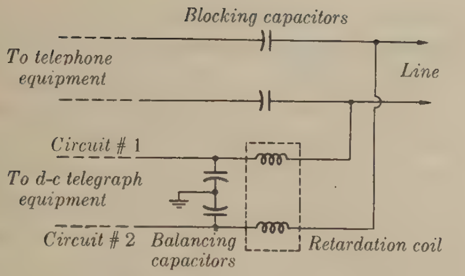


FIG. 18-24 Circuit for compositing telephone and telegraph signals.

Telephone communication takes place by means of alternating currents at voice frequencies coupled into the line through the transformers. The telegraph signals consist of d-c impulses impressed upon the line simultaneously with the telephone signals. With reasonably good impedance balance between the two wires of the pair, the telegraph current divides approximately equally between the two wires and there is little interference with the telephone transmission. The telephone currents do not introduce any material voltage into the telegraph circuit since they are transmitted as a-c signals on a two-wire system.

B. Composited Circuits

The compositing method for deriving d-c telegraph circuits from wires used for telephone service is based on frequency discrimination, or filtering which is a method of selecting the desired electrical frequencies and rejecting others. A *composite* set, shown schematically in Figure 18-24, consists of a retardation coil and capacitors, and provides a low-pass filter (approximately 0-80 c/s) for telegraph currents and a high-pass filter for the telephone currents. The compositing method has the advantage of providing two telegraph channels from a pair of wires used for telephone service. Compositing is usually applied to open wires in order to derive as many d-c facilities as possible. Generally, the effect of leakage is less on composited circuits than it is on simplexed circuits.

18.6 Telegraph transmission principles. In telegraph transmission, we are concerned with the reproduction of the transmitted telegraph message at the receiving end at a satisfactory rate, without error and without interference to other services. The characteristics of the circuits are such that the transmitted signals are not always reproduced accurately at the receiving end. Telegraph signal transmission is accomplished on a two-current basis, that is, by transmitting impulses of steady current in one direction (marks) interspersed by intervals of no current (spaces) in the case of neutral operation and by transmitting impulses of current in one direction interspersed by reversals of current in the case of polar operation. In either neutral or polar operation, there is a change from one current condition to another, that is, from mark to space or vice versa and this change is known as the *transition*. The change of current from the marking to the spacing condition, or vice versa, can be plotted with respect to time and a waveshape diagram produced. These waveshapes are an important aid in the study of telegraph transmission.

A. Waveshapes in Neutral Telegraph Systems

In any neutral telegraph circuit, if the time required for the direct current to establish itself and to decay could be ignored, the waveshape for the letter *A* in the Morse Code would be as illustrated in Fig. 18-25 (a). In practice, every telegraph circuit has some series inductance; each line relay adds some inductance and, in the case of composited circuits, each retardation coil adds more inductance. The waveshape with series inductance present is represented by Fig. 18-25 (b). Each current pulse has a sloping curve from zero to maximum at

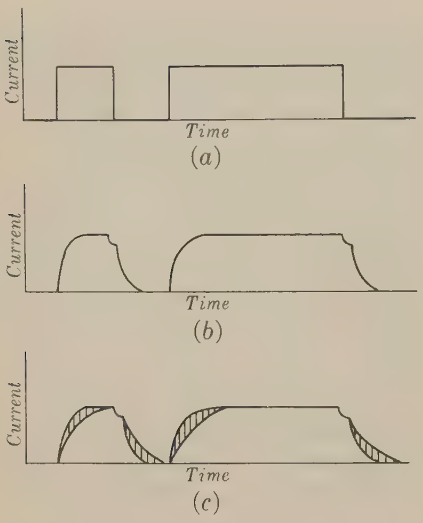


Fig.18-25 Wave shapes in neutral telegraph operation.

the closing of the key and from the maximum value to the point where the arc is broken at the opening of the key. If, in addition, the effect of capacitors is considered, there is a further sloping of the pulse as shown in Fig. 18-25 (c). The shaded portion represents the effect of the circuit capacitors over and above the effect of the inductances. Telegraph relays normally have two windings, an operating winding which, when energized, closes the relay and a bias winding which holds the relay open in the absence of current in the operating winding. The value of current in the bias winding is usually half that of the line current which flows through the operating winding. Thus, when the key is closed, the value of the current in the operating winding rises to a point where it overcomes the effect of the bias winding and moves the relay armature from the space to the mark position. Similarly, when the key is opened, the current in the operating winding must fall below the bias current before the relay armature moves back to the space position. The relay operating points on a waveshape occur during the transition periods, known as the *space-to-mark* (abbreviated S-M) *transition* and the *mark-to-space* (abbreviated M-S) *transition.* At the instant the key is closed, the line current starts to rise in the receiving relay but does not reach the relay operating point until a few milliseconds later. This means that there is a delay between the closing of the sending key on an S-M transition and the operation of the receiving relay This is called the *space-to-mark transition delay* and is abbreviated S-MTD. In a similar manner, when the sending key is opened, the line current in the receiving relay does not become zero instantaneously and the armature is held on its marking contact until the line current falls to a value below the bias current. This time delay is called the *mark-to-space transition delay* and is abbreviated M-STD. The magnitude of these delays ranges from a fraction of a millisecond to several milliseconds; the occurrence of transition time delays is illustrated diagramatically in Fig. 18-26.

The S-MTD and the M-STD are determined entirely by the characteristics of the circuit and, though the two delays may not be equal, each transition delay is always a constant for any given circuit under any given set of adjustments. Each mark, regardless of its length, must start with an S-M transition and end with an M-S transition. The S-MTD cuts off the beginning of each mark impulse and the M-STD adds to the end of each mark impulse. If the two delays are equal, the length of the mark impulse

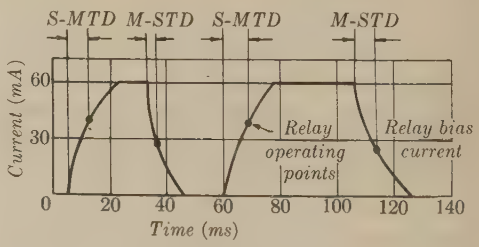


Fig.18-26 Transition delays.

is unchanged by transmission over the circuit. Each space, regardless of its length, begins with an M-S transition and ends with an S-M transition. The M-STD cuts off the beginning of each space and the S-MTD adds to the end of each space. Thus, the S-MTD and the M-STD have opposite effects on the space and the mark. If the two delays are equal, the length of each space will be unchanged by transmission over the circuit. The transmission is considered to be perfect if the received marks and spaces are exactly the same length as the transmitted marks and spaces; otherwise, the transmitted signal is said to be distorted.

B. Bias Distortion

The requirement for perfect transmission is that the S-MTD be equal to the M-STD. If the M-STD is greater than the S-MTD, all marks will be lengthened and all spaces will be shortened. This condition is known as *marking* or *positive bias.* If the S-MTD is greater than the M-STD, all marks will be shortened and all spaces will be lengthened. This condition is known as *spacing* or *negative bias.* Since the S-MTD and the M-STD are measured in milliseconds, the difference between the two is often referred to as *millisecond bias.* This bias will be constant for any given circuit and is given by the formula

millisecond bias = M-STD − S-MTD

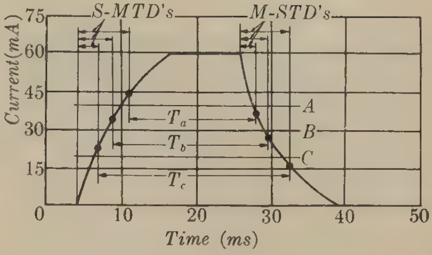


FIG. 18-27 Effect of relay bias current on signal length.

The effect on transmission of a given millisecond bias varies with the lengths of the marks and spaces transmitted. As an example, consider a manual telegraph circuit where the dashes (long marks) are three times as long as the dots (short marks). As the speed of transmission increases, the lengths of the dots and dashes decreases. With a slow transmission speed, dots 30 msec. long and dashes 90 msec. long may be assumed. If the circuit has a positive millisecond bias of 10 msec., the dots will be 40 msec. long and the dashes 100 msec. long at the receiving end. The signal will still be readable since the 3:1 ratio has been approximately maintained. Under conditions of fast transmission, where the dots are 5 msec. long and the dashes are 15 msec. long, the same positive millisecond bias of 10 msec. will make the dots 15 msec. long and the dashes 25 msec. long. Difficulty will now be experienced in reading the signal since the dashes are now less than twice as long as the dots. The waveshape of a typical mark impulse in neutral operation with a line current of $60mA$ is shown in Fig. 18-27. The horizontal lines A, B and C represent different values of relay bias currents. The relay operating and releasing points (designated by heavy dots) are indicated for these three values of bias current. When the normal bias current of $30mA$ (line B) is used, the length of the mark signal is that indicated by T_b. It is seen that increasing the relay bias current increases the S-MTD and decreases the M-STD thus producing spacing bias. A reverse condition results from lowering the relay bias current and gives rise to marking bias. The same effect as that obtained by raising the bias current, which shifted the relay operating points to the narrow part of the wave, is obtained by decreasing the line current while holding the bias current constant. Increasing the line current, which effectively shifts the relay operating points to the broader part of the wave, gives rise to the same condition as decreasing the bias current. Thus, a decrease in line current in a neutral system produces spacing bias while an increase in line current produces marking bias.

C. Waveshapes in Polar Telegraph Systems

In a polar circuit, the sending relay connects a negative voltage to the line for the marking condition and a positive voltage to the line for the spacing condition. The resistance at the sending end is adjusted so that the line current has a steady state value of $+35mA$ for marking and $-35mA$ for spacing. In polar circuits, as in other circuits, the change of the line current from marking to spacing and from spacing to marking is delayed because of the line inductance and capacitance and the capacitance between the line and ground. The M-S transition and the S-M transition are gradual changes as shown in Fig. 18-28. The fact that the M-S and the S-M waveshapes are identical in form is a valuable feature of polar operation. However, to obtain full advantage of this feature, the relay operating points must be symmetrically located with respect to the zero current line on the waveshape. If the relay operating points are the same distance on each side of the zero current line, the S-MTD and the M-STD will be equal and there will be no bias in the received signal. Thus, undistorted polar transmission depends on the following three conditions:

(i) that equal and opposite potentials be applied at the sending end of the circuit

(ii) that the resistance of the circuit remain constant for both positions of the sending relay armature

(iii) that the operating points of the relay be located symmetrically about the zero current line of the waveshape diagram in order that equal transitions will be secured.

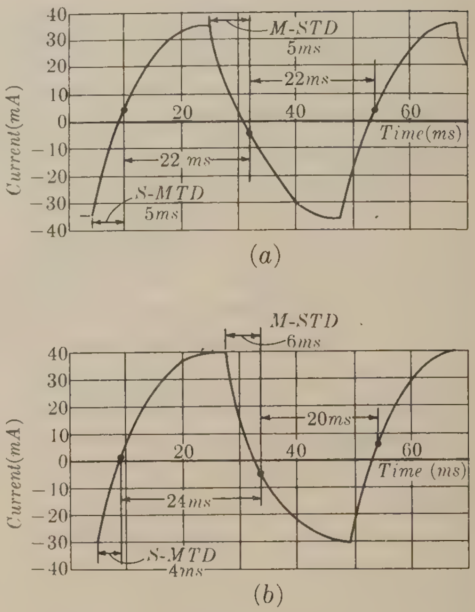


FIG. 18-28 Signal wave shapes in polar telegraph circuits (a) Effect of equal line currents and symmetrically located relay operating points (b) Effect on signal lengths of unequal polar line currents.

D. Characteristic Distortion

Thus far, a transition has been assumed to start when the line current is at the steady-state marking or spacing condition. However, there are situations where the start of the transition does not occur when the line current is at its steady-state value. In practice, the time required for the current to change from one steady-state condition to the other is sometimes greater than the minimum time interval between the transitions in the signals. Some transitions must then occur while the line current is in the process of changing from the previous transition. Such conditions are shown diagramatically in Fig. 18-29. These transitions have a different delay time from the transitions starting when the

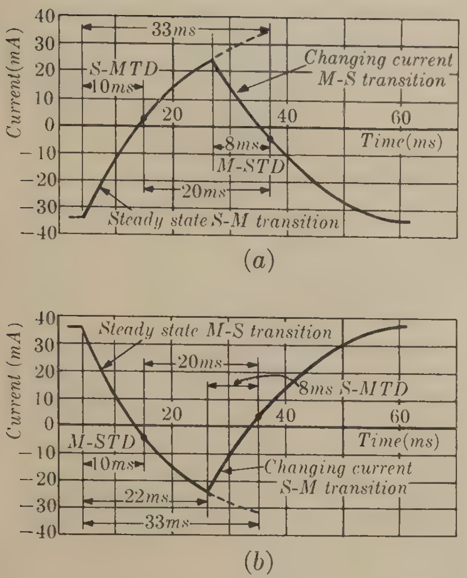


FIG.18-29 Changing current transitions (a) Changing current mark-space transition (b) Changing current space-mark transition.

line current is in the steady state condition and must therefore be distinguished from them. The amount of *changing current transition* delay is dependent upon the value of the line current at the start of the transition. The value of the line current is, in turn, dependent upon the time interval between the changing current transition under discussion and the previous transition which started the line current to change. Since the time interval between these two transitions is equal to the length of the sent impulse, it is the impulse length which finally determines the transition delay under a given set of conditions. Thus, the lengths of the received signal impulses are affected by the presence of changing current transitions. This phenomenon is known as *characteristic distortion*. Fig. 18-30 shows characteristic distortion effects at various speeds of operation.

The magnitude of the characteristic distortion is inversely proportional to the length of the sent impulses and the nature of the distortion is to shorten the received short impulses. An opposite effect is also possible. The characteristics of the circuit may be such that the line current tends to increase momentarily at the completion of each transition to a value greater than the steady state owing to transient effects. If the next transition occurs at such an instant, the transition delay is greater than the delay on the preceding transition; this causes the received mark or space signal to be lengthened. This effect is called *positive characteristic distortion*. However, as the transient effect causes the line current to oscillate around the steady state, it is possible that the next transition might occur when the line current is below its steady state value. Thus, transient conditions may cause either positive or *negative characteristic distortion*. Positive characteristic distortion is not often encountered, however.

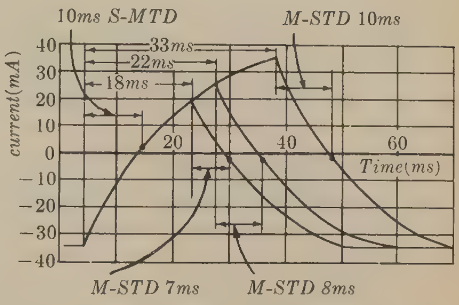


FIG.18-30 Effect of characteristic distortion on signals of various durations.

E. Comparison of Characteristic and Bias Distortions in Polar Circuits

The comparison of characteristic distortion and bias distortion in polar circuits follows.

(i) Characteristic distortion is related to the amount and arrangement of the capacitance, inductance and resistance of a circuit. Except in neutral systems, these factors do not affect the amount of bias distortion.

(ii) Bias distortion is caused by unequal marking and spacing line currents and biased relays; these conditions do not affect characteristic distortion.

(iii) Characteristic distortion, because it is caused by the capacitance, inductance and resistance of a circuit, which, except for resistance, are unchanging in value, varies only a small amount from day to day on a given circuit. Bias distortion, because it is caused by unbalanced voltages, ground potentials and relays out of adjustment, may vary from hour to hour on a circuit.

(iv) The effect of characteristic distortion depends upon the length of the impulses transmitted. The effect of bias distortion is independent of the length of the impulses.

(v) For a given length of impulse, the effect of characteristic distortion is independent of whether it is a marking or a spacing impulse. The effect of bias distortion is always opposite on a marking impulse to what it is on a space.

F. Fortuitous Distortion

The form of distortion caused by such factors as powerline induction, momentary battery fluctuations and other sporadic effects which displace miscellaneous received transitions by various amounts intermittently is known as *fortuitous distortion.* At times, this effect may be large enough to produce a complete failure of the circuit. In the transmission of miscellaneous signals, the combined effect of characteristic and fortuitous distortion on the received transitions is sometimes known as *jitters.*

18.7 Loading. One of the major problems encountered in communication by land lines is that of transmitting signal currents over long distances without excessive attenuation or distortion. These effects are often the deciding factor in determining whether or not the signal can be accurately reproduced at the receiving terminals. The total attenuation at a given frequency depends on

(i) the length of the circuit

(ii) the attenuation per unit length

(iii) the energy transfer at the two ends of the circuit and at any junctions of dissimilar sections

(iv) energy losses of the apparatus that may be associated with the circuit.

If the circuit is long enough, the total loss tends to become so great that the energy reaching the distant end is insufficient to operate the terminal apparatus. It should also be remembered that the transmission of the required volume of energy is not the only consideration. Currents of different frequencies are attenuated unequally as they pass along the circuit and distortion results. Thus, there is a complex relation between the various physical characteristics of a land line which determines its efficiency for telegraph or telephone transmission. The application of *line loading* is widely used in solving these problems of long distance wire communication. Line loading makes certain improvements in the transmission efficiency of the circuit through one or more of the following effects:

(i) a reduction of the attenuation per unit length of the circuit

(ii) a more even attenuation of the various frequencies within the band of frequencies to be transmitted, thereby reducing distortion

(iii) a more nearly constant characteristic impedance for the frequencies within the band to be transmitted, a consideration which is most important in the satisfactory operation of repeaters but also of some importance in considerations having to do with the termination of the circuit.

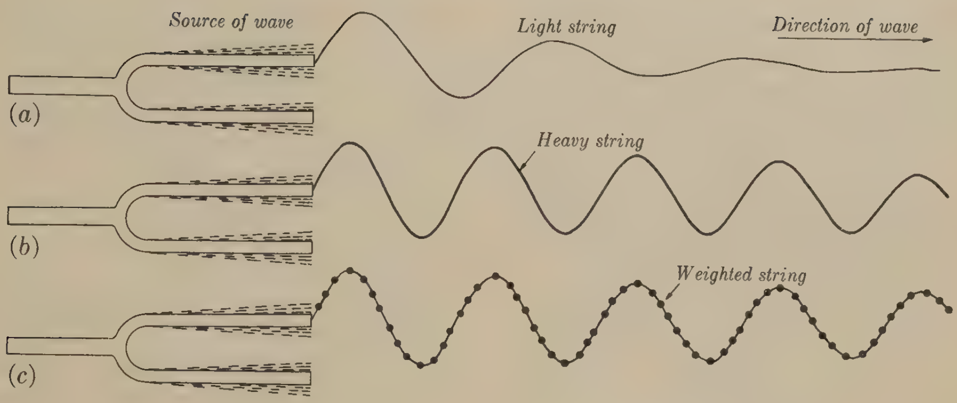


FIG. 18-31 Mechanical analogy to illustrate the principles of loading.

A. Mechanical Analogy of Loading

Telephone and telegraph transmission deals with the propagation of electric waves along transmission lines. While the problems associated with transmission lines are adequately explained by mathematical analysis, the physical concepts of wave propagation and loading are also helpful in understanding these problems.

Fig. 18-31 (a) shows a mechanical wave motion being propagated along a very long string, the energy being supplied by the vibration of a tuning fork. As the wave proceeds along the string, the amplitude decreases and the wave dies out. The dissipation of energy along the string is called *attenuation*. If a heavier string is substituted, as indicated in Fig. 18-31 (b), the wave is propagated farther along the string or is less attenuated. If it is undesirable to increase the weight of the string uniformly along its length, the same effect may be obtained by placing many small weights, equally spaced, along the string as shown in Fig. 18-31 (c). It is found that increasing the effective weight of the line reduces the wavelength and velocity of propagation.

Further experimenting with tuning forks of various frequencies shows that there is a minimum number of these tiny weights required in each wavelength of line. If the frequency of the wave propagation is increased, the wavelength is decreased and the number of weights per unit length of line must also be increased. For a given number of weights per unit length of line, there is a critical upper frequency above which transmission will not take place.

Adding weights to a string to increase its mechanical inertia is analogous to adding series inductance to a transmission line to increase its electrical inertia. The weights oppose sudden changes in motion of the line while the inductance opposes sudden changes in current. Loading coils are inserted at regular intervals along long-distance telephone and telegraph lines to reduce attenuation and extend the useful length of the line. From the behaviour of the loaded string, it may be concluded that, on a loaded circuit, the loading coils must be spaced so that there are several in each wavelength of any frequency that it is desired to transmit. Thus, loading suitable for a transmission of 1000 c/s is not necessarily suitable for a transmission of 3000 c/s since the wavelength at 3000 c/s will be considerably shorter than at 1000 c/s. Accordingly, the number of loads per wavelength is reduced and the load coil, instead of aiding the transmission, hinders it just as scattered heavy weights would hinder the transmission of the mechanical wave along the string. The velocity of the waves must be large enough so that each wavelength passes over approximately nine coils. If the proper number of coils is not encountered, the losses increase rapidly and cut-off follows.

B. Loading Coil Design and Practices

In the design and manufacture of the various types of loading coils used on long distance lines, there are a number of requirements other than merely providing a specified inductance value. Loading is effective in reducing attenuation only when the increase in the alternating-current resistance of the circuit is held within certain limits. A loading coil should, therefore, have minimum resistance and minimum losses in its core. The inductance of each coil must be accurately divided so that one half of the inductance is inserted in each wire of the circuit to maintain circuit balance. This requirement is very exacting and unless the two windings of the coil are identical in every respect, noise and crosstalk will result. Fig. 18-32 illustrates a method of winding coils to give a high degree of balance. Loading coil cores are generally toroidal in shape and made of a low-loss magnetic material such as permalloy or molybdenum-permalloy. The magnetic material is first powdered, then mixed with shellac to insulate each particle and finally pressed into solid rings. This process gives the coils a high degree of magnetic stability. The coils are mounted in welded steel pots of standard sizes. The following considerations are fundamental to good loading practice.

(i) In connecting the loading coil, care must be exercised to prevent reversal of one winding thereby neutralizing the inductance of the coil.

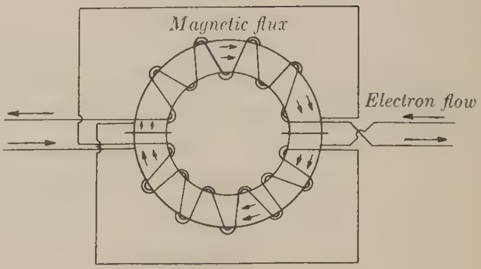


Fig. 18-32 Winding of a single loading coil.

(ii) The inductance values of loading coils should not vary by more than 2% where the circuit is used in connection with telephone repeaters.

(iii) The loading coil spacing should be accurate to within 2% where the circuit is used in connection with telephone repeaters.

(iv) To prevent loading coil magnetization, the line current used for telegraph operation should not exceed the specified limits for the particular type of loading.

C. Coil Spacing

Another important aspect of loading is the spacing of the coils. These coils must be close enough together so that the effect of distributed inductance will be obtained. Capacitive reactance decreases and inductive reactance increases with an increase in frequency; lumping the inductance tends to increase this effect. On a line with uniformly distributed constants, current passes from the positive wire to the negative wire through the small capacitive elements distributed along the wire, whereas, in the simulated T-network this passage of current can take place only in the middle of the T-sections. Thus, the inductive effect is greater than in the uniform line since somewhat larger currents flow through the inductances and the capacitive effects are correspondingly less. The greater the number of sections, the less is the difference in behaviour between the uniform line and the network. Since the difference is increased with higher frequency, the coil spacing on loaded circuits is so chosen that the change produced by frequency over a given range does not differ greatly from the effect that would be produced on a continuously loaded circuit of the same average line constants. The value of the loading inductance is thus a factor in determining the spacing, or, for any chosen standard spacing, the distance between coils is a factor in determining the inductance values of the coils.

D. Critical Frequency

A band of frequencies between 200 and 2700 c/s will transmit telephone conversations with adequate fidelity and intelligibility. Where a uniformly loaded circuit would have an impedance and attenuation which might vary little with change in frequency, the practical loaded circuit, i.e. the simulated network, has an impedance and an attenuation constant which increase with frequency near the upper limit of the voice-frequency band. The attenuation rises so rapidly that, a few hundred c/s above the upper limit, the amount of current passed is negligible and the circuit is said to be *cut off*. The critical frequency at which cut-off occurs is determined by the formula

$$f_c = \frac{10^6}{\pi\sqrt{L_o C_o}}$$

where f_c = critical upper frequency in cycles per second
L_o = inductance of equivalent network section in henries
C_o = capacitance of equivalent network section in farads.

The effects of loading on attenuation and distortion are not easily explained in terms of physical concepts. Mathematical formulas for the impedance and attenuation of transmission lines are given in the applied mathematics section.

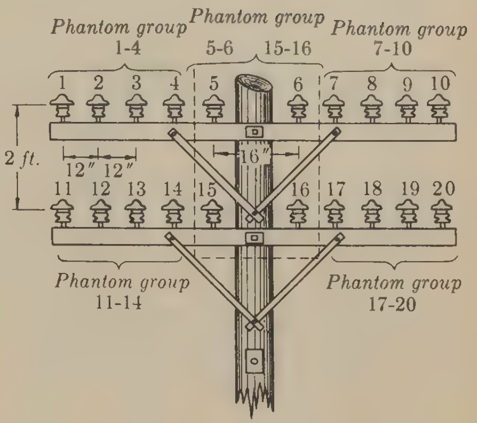


FIG. 18-33 Wire configuration for open-wire line carrying voice-frequency side and phantom circuits.

18.8 Circuit facilities. Wire facilities used for carrying telephone and telegraph traffic may be classified in several ways according to their uses, or on the basis of their physical and electrical peculiarities. Thus, facilities may be classified as toll or local exchange, or they may be classified as those transmitting voice frequencies and those transmitting carrier frequencies, or again, as open-wire and cable circuits.

A. Open Wire

The majority of long-distance telegraph facilities and some toll telephone services are carried on open-wire circuits. Prior to the development of satisfactory repeaters, open-wire circuits were necessary for long-distance communication in order to keep down the attenuation. Open-wire circuits are carried on cross-arms as shown in Fig. 18-33. In this illustration, each cross-arm carries ten wires which are numbered consecutively starting with the left hand pin on the top cross-arm. The standard wire layout on two cross-arms provides for ten *side circuits* and five *phantom circuits*. A phantom circuit is a method of providing an additional circuit over two pairs of wires and is applicable to both open-wire

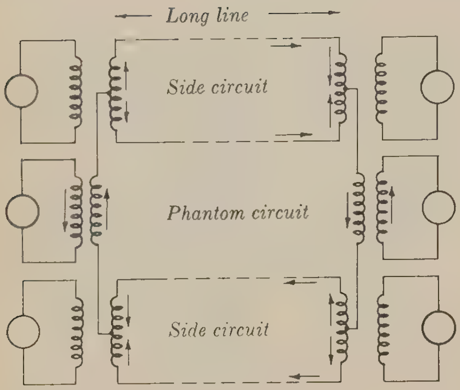


FIG.18-34 Derivation of a phantom circuit.

and cable circuits. The provision of a phantom circuit is made possible by the use of a special type of transformer known as a *phantom repeating coil.* In Fig. 18-34 the transformers are tapped exactly at their electrical centres and, if the impedance of each line is the same, the currents from the phantom circuit divide equally in the coils. Since these currents flow in opposite directions through the windings, their magnetomotive forces neutralize and no flux is produced to induce currents in the side circuits. Instantaneous directions of current flow for a phantom circuit are shown in Fig. 18-34. The side-circuit currents flow the same as if the phantom were not installed. In Fig. 18-33, phantoms are derived from wires 1-4, 7-10, 11-14, 17-20 and 5-6, 15-16. This last group of four wires is called a *vertical* or *pole-pair* phantom and has somewhat different circuit characteristics because of the different spacing and configuration of the wires.

Long distance, open-wire circuits are not loaded because the electrical characteristics of these circuits change sc markedly with varying weather conditions. In dry weather, open-wire loading is effective in reducing the attenuation but due to the increased leakage in wet weather, loading actually increases the attenuation. Therefore, in order to maintain overall transmission stability, the lines are not loaded and the attenuation is compensated for by the use of additional repeaters.

B. Cable

Almost all local exchange circuits, a high percentage of toll telephone circuits and some telegraph circuits are today carried in cable. Cable is composed of copper conductors insulated with a spiral wrapping of dry paper enclosed in a lead sheath. A method has been developed and has in recent years been extensively used for insulating the conductors by paper pulp. This has resulted in the development of cables capable of carrying a very large number of circuit pairs. The largest cable in use in congested city areas contains 2121 pairs or 4242 conductors. Additional protection such as jute covering and armoured sheathing is provided for cables in underground, submarine or exposed locations.

Coaxial cables are now extensively used for telephone toll circuits. By employing carrier techniques, a single coaxial cable may be used to transmit several hundred telephone conversations simultaneously or a single standard television program. The attenuation losses in coaxial cables are large at the very high carrier frequencies at which they are used. As a result, the application of coaxial line facilities involves the use of high-gain amplifiers spaced at very close intervals along the line. As a result, the installation cost of coaxial line facilities is very high but once they are installed, the operating costs are considerably lower than for conventional facilities.

C. Networks and Filters

In the operation of telephone and telegraph circuits, it is frequently necessary to reduce the voltages and currents

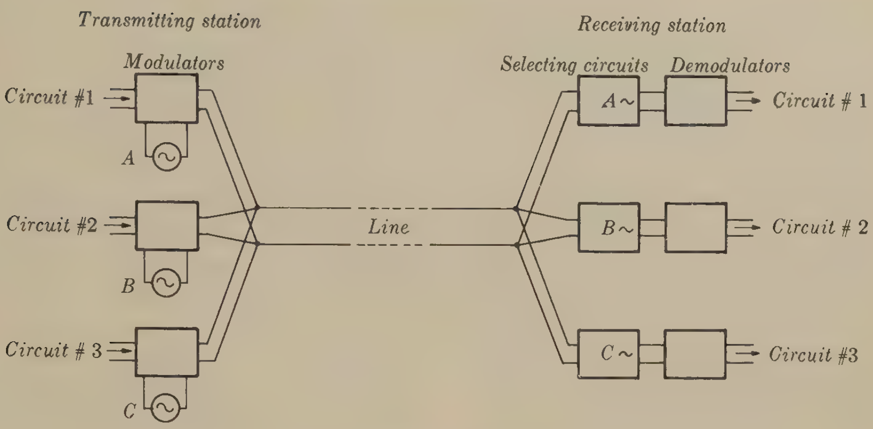


FIG. 18-35 Principles of carrier operation.

at various points within the circuits. To accomplish this, attenuators and pads are inserted; *T* and π pads are most commonly used for this purpose. One of the factors tending to decrease the intelligibility of telephone conversations is the unequal attenuation of currents of different frequencies as they pass over the circuit. Networks known as *attenuator equalizers* are used to compensate for this unequal attenuation. These equalizers are usually associated with the repeaters in the circuit. An equalizer network consists of retardation coils, capacitors and resistors so proportioned and arranged that their frequency attenuation characteristics are complementary to the line characteristics that produce the distortion. Other equalizers are used to reduce phase distortion along the line. In telephone and telegraph work, it is sometimes desirable to suppress or eliminate currents of undesired frequencies and this is accomplished by the use of filters. On high grade circuits, crystal filters are often used.

18.9 Carrier systems. The carrier principle is used in both telephone and telegraph communication to provide independent transmission of several messages over a single circuit at one time. As in the case of wireless telegraphy and telephony, carrier transmission over wire circuits is accomplished by impressing low-frequency signals on a high-frequency carrier. Several carriers operating at different frequencies and carrying separate signals may be sent over a single wire circuit at one time. At the receiving end, the carriers are separated from each other by frequency filters and the intelligence impressed on each is recovered in a separate channel. Carrier transmission is accomplished by

(i) providing by means of vacuum-valve oscillators, or by other means, the currents of different selected frequencies to be used as carriers
(ii) impressing upon each carrier by a modulation process, the message current from the terminal telephone or telegraph station
(iii) separating or selecting the several modulated carrier currents at the receiving end by means of filters
(iv) separating or restoring from the selected carrier current, the original message current for transmission to the receiving terminal telephone or telegraph station. This is a demodulation process.

Fig. 18-35 illustrates the general arrangement required for transmission in a single direction. If it is desired to

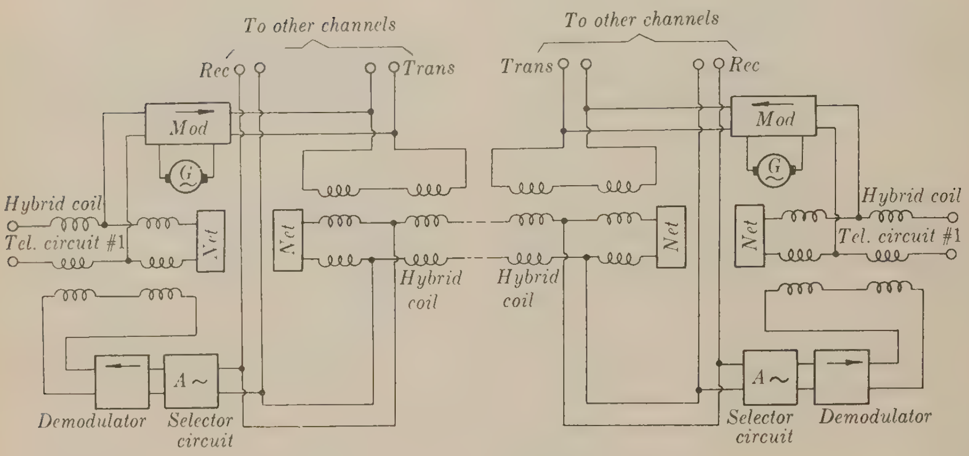


FIG. 18-36 Method of obtaining two-way carrier operation.

transmit in both directions, hybrid coils must be inserted between the terminal apparatus and the carrier line and between the terminal apparatus and the message line as shown in Fig. 18-36. This method, however, is rarely used because of the inherent balance and crosstalk difficulties. Practically all modern carrier systems operate on a four-wire basis with separate carrier channels for transmission in the two directions for each circuit. Carrier techniques provide both economy of circuits and higher quality transmission.

A. *Low-Frequency Carrier Systems*

Carrier systems which fall into this category utilize carrier frequencies up to about 30 Kc/s. The two main types of systems included in this group are the Type *B* telegraph system and the Type *C* telephone system.

The Type *B* telegraph carrier system provides for imposing ten two-way telegraph channels on an open-wire telephone circuit which may also be composited for ordinary grounded telegraph operation. Twenty carrier frequencies, divided into two groups, are employed; the lower ten for transmission in one direction, the upper ten for transmission in the other direction. The range of frequencies used extends from about 3 to 12 Kc/s. The separation between carrier frequencies ranges from slightly less than 200 c/s at the lower end of the range to nearly 1000 c/s at the upper end. With the carrier frequencies grouped for transmission in the two directions, separation at terminal and repeater points is secured in each case by a pair of directional filters, one of which passes only frequencies below 6000 c/s and the other, only frequencies above that value. Since only the single frequency of the carrier itself is transmitted over the line, the channel filters need only be simple tuned circuits and the modulator and demodulator circuits may also be relatively simple.

A second type of telegraph carrier system is called the *voice-frequency system* because the carrier frequencies utilized are all within the normal voice-frequency band. This system is not essentially different from the Type *B* system except for the lower frequencies employed. As the transmitting medium in this system is either actually or effectively a four-wire circuit, there is no problem of separating the transmitting and receiving channels; this allows the same carrier frequencies to be used for transmission in both directions. This system provides 12 two-way telegraph circuits, employing carrier frequencies which are multiples of 85 c/s between 425 c/s and 2295 c/s, with an adjacent-channel separation of 170 c/s.

The Type *C* telephone carrier system is designed for long distance circuits and operates on open-wire lines. It provides three telephone circuits in addition to the normal voice-frequency circuit. The carrier frequency is suppressed in the modulator and only one sideband is transmitted over each channel. A 3000 c/s separation is maintained between adjacent channels to permit transmission of a sideband about 2500 c/s wide. Directional filters are used at terminal and repeater points to allow the use of a single amplifier for all three channels. One thousand cycles per second signalling is employed and this ringing current is transmitted over the system in the same manner as the normal voice-frequency currents. In the more recent modifications of the Type *C* system, push-pull amplifiers at repeater points have been replaced by negative feedback amplifiers and *varistors* (copper-oxide rectifier modulator units) are used in lieu of vacuum-valve modulators.

B. Broad-Band Carrier Systems

Broad-band carrier systems differ from low-frequency systems in that they operate through a much wider band of frequencies and are capable of providing more telephone or telegraph channels per system. The Type *J* system for open-wire lines employs a frequency range from about 36 Kc/s to 140 Kc/s and provides 12 telephone channels over a single pair. Since a single pair of wires may also carry, in addition to the regular voice-frequency channel, the three channels of the Type *C* system, it is possible to have as many as 16 telephone channels on a single pair. In general, broad-band carrier systems have higher attenuation losses and are more susceptible to noise and crosstalk than low-frequency systems.

The Type *J* system does not differ greatly from the Type *C* system except that more than one stage of modulation and demodulation are necessary in order to permit the use of the most desirable group of frequencies for the initial channel modulation while at the same time applying to the line a band of frequencies most suitable for transmission. The lowest of the 12 carrier channels uses a frequency of 64 Kc/s and an adjacent channel separation of 4000 c/s is employed. This group of carrier channel frequencies has been selected primarily because high-grade crystal filters can be most economically built for operation in this general range. Also, it is important that the range be high enough so that the lowest harmonic of the lowest frequency is above the highest frequency in the band. Thus, the second harmonic of 60 Kc/s, the lowest frequency in the lower sideband of the 64 Kc/s channel, is 120 Kc/s, which is well above the highest frequency in the upper sideband of the highest carrier channel used, 108 Kc/s. General design and manufacturing economy is achieved by using this same group of carrier channel frequencies for all broad-band systems.

After the twelve channel carriers have been modulated with the voice frequencies and all but the lower sidebands have been eliminated by means of appropriate band filters, the entire group of resulting frequencies, 48 Kc/s in width, is translated by an additional modulation process to the band of frequencies it is desired to transmit over the line. The range of this final band depends upon the direction of transmission since transmission in both directions is carried out over a single pair of wires. Transmission in one direction occupies the band between 36 Kc/s and 84 Kc/s and in the other direction, between 92 Kc/s and 140 Kc/s. A bandwidth of more than 3000 c/s is transmitted; this is a distinct improvement over the 2500 c/s band possible with the Type *C* system. Since the line losses at the high frequencies are relatively higher, the Type *J* system requires the use of repeaters at considerably closer spacings than used in the Type *C* system.

The Type *K* carrier system for use with cable facilities, utilizes the same types of crystal filters, varistors and equalizers as the Type *J* system but, as far as transmission over the line is concerned, the techniques employed in the two systems are quite different. Since any practicable method of cable circuit loading causes transmission to cut off at a relatively low frequency, non-loaded cables are used for Type *K* broad-band carrier transmission. Because the attenuation of non-loaded conductors is very high and increases with frequency, it is desirable to keep the maximum frequency to the lowest possible value. Accordingly, the band of frequencies selected for transmission in the Type *K* system is that between 12 Kc/s and 60 Kc/s. By using pairs in separate cables for transmission in the two directions, only one 48 Kc/s band is required. High-gain repeaters must be used at spacings of about 16 miles, as compared to average spacings of 80 miles used in the Type *J* system and 150 miles used in the Type *C* system.

Broad-band carrier systems using the same initial twelve channels as used in the Type *J* and Type *K* systems have been adapted for use with coaxial cable circuits. By spacing high-gain repeaters no more than five miles apart, it is possible to obtain as many as 500 telephone channels from a pair of coaxial cables. Coaxial cable circuits are also used to carry television programs between cities. Because of the relative complexity of the terminal equipment and the large number of repeaters required, coaxial cable circuits have a very much higher installation cost. In recent years, there has been a trend toward the use of microwave radio channels for the purpose of telephone multiplex and the cost and reliability of such systems compete favourably with coaxial cable methods; the carrier techniques used in microwave radio are similar to those used in coaxial land lines.

18.10 Noise and crosstalk. The intelligibility of wire communication depends on the absence of noise and crosstalk. If each communication circuit were completely isolated from all other electrical circuits of every kind, no potentials, other than those deliberately introduced for the purpose of the transmission of information, would be present. In practice, nearly every long-wire communication circuit is in close proximity with other communication circuits as well as power lines. Therefore, it is necessary that communication circuits not only be efficient in transmitting electrical energy without distortion and too much attenuation but also that they be protected against induced electrical currents coming from adjacent communication circuits and other electrical circuits. Any two long paralleling circuits that are not balanced by means of transpositions or otherwise, may introduce crosstalk to such a degree as to interfere with their practical use. Furthermore, since crosstalk is an inductive effect, its magnitude tends to increase with the length of the paralleling circuits, the energy level and the frequency of the transmitted currents. Thus, the use of repeaters and carrier systems increases the possibilities of crosstalk. Crosstalk may be produced by both magnetic and electric induction between circuit pairs.

A. Crosstalk Reduction

There are several methods of eliminating or, at least, partially reducing crosstalk between parallel circuits. One possibility is to arrange the paralleling wires in such a configuration that the effect of the magnetic and electric fields of one pair will be the same at both wires of the other pair, thus leaving no residual difference to cause circulating currents in the second pair. Usually such wire configurations are not practicable. Another partial solution is to reduce the separation between the wires of either or both pairs and, if practicable, increase the separation between the pairs themselves. In practice, it is possible to take advantage of this method to only a limited extent. In cable circuits, the two wires of each pair are close together but so are the pairs; in open-wire lines, there is a considerable separation between pairs but the two wires in a pair cannot be placed close together.

B. Transposition

A third alternative in the reduction of crosstalk is the use of transpositions. The principle involved can best be understood by reference to Fig. 18-37 which represents four wires in a typical open-wire configuration. The relative pin positions of wires 1 and 2 are interchanged in the middle of the short section *S*. Under these conditions, it is evident that when equal and opposite currents are flowing in the two wires of circuit *A* the voltages induced in wires 3 and 4 will be in opposite directions on the two sides of the point where the circuit *A* is transposed. Thus, as indicated in the illustration, while e_3 is larger than e_4 and e'_3 is larger than e'_4, e_3 is exactly equal and opposite to e'_3 and e_4 is exactly equal and opposite to e'_4. There is therefore no net voltage induced in either wire 3 or wire 4 and consequently no crosstalk in circuit *B*. The same net effect could be obtained by inserting the transposition in the disturbed circuit *B*, leaving the wires in the disturbing circuit running straight through. Thus, it is seen that the transposition of a single pair of wires of two interfering circuits eliminates crosstalk between both pairs. Transposition of both pairs at the same point would nullify the attempt to reduce crosstalk.

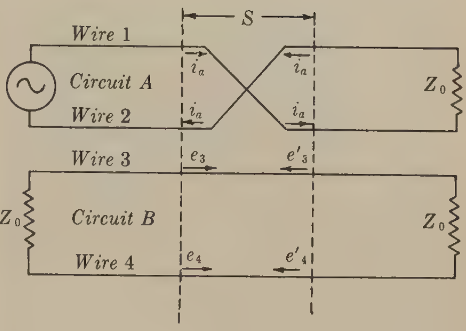


FIG. 18-37 Principle of transposition.

While a single transposition is effective in eliminating crosstalk in a short section *S*, it would not be sufficient to reduce crosstalk in the whole line. Because of the attenuation, the current and voltage near the energized end of the disturbing circuit are much greater than the current and voltage near the other end. Thus, it can not be expected that the induced crosstalk on the energized side of the transposition would be neutralized by the weaker crosstalk induced on the other side of the transposition. Also, in a long circuit, several wavelengths may be included in the propagation of a voice current from one end to the other. Since crosstalk is an induced effect, its instantaneous value in any small section *S* depends on the position of *S* with respect to the cycle of current in the disturbing section. If *S* is so located that the current or voltage in it has a maximum value, either positive or negative, it can not be expected that the crosstalk induced in this section will be neutralized by the crosstalk in a similar section, located at a point in the line where the current or voltage has a value of nearly zero at the same instant. Thus, it is necessary that transpositions be installed at frequent intervals with respect to the wavelength of the propagated current.

For voice-frequency transmissions, where frequencies are relatively low, it is not difficult to achieve a sufficiently good approximation to this condition because of the relatively long wavelengths. Where high-frequency carrier systems are used, the wavelengths are so short as to require very closely spaced transpositions. In open-wire lines, spacings as close as every second pole are used in practice where the wires are carrying frequencies up to 140 Kc/s. The degree of effectiveness of such closely spaced transpositions is dependent on the accurate spacing of the poles and is further limited in practice by economic rather than theoretical factors.

The most striking feature of cable circuits with respect to crosstalk is that the conductors are crowded very closely together. This close spacing of the two wires of a pair in which equal and opposite currents are flowing tends to minimize the external effect of the electromagnetic field of the pair. In the process of manufacture, the cable conductors are very thoroughly transposed by twisting the two wires of each pair together, by twisting the two pairs of each group together to form *quads* and by spiralling the quads in opposite directions about the cable core. Cables are also manufactured and installed so that their conductors are practically free from series resistance unbalances and insulation leakages. However, the close spacing of many circuits within the cable sheath, as well as their proximity to the sheath, tends to offset these advantages to a considerable extent. At voice frequencies, the magnetic induction between circuits in a cable is usually so small as to cause relatively little crosstalk. However, the capacitive unbalances usually remain large enough to cause objectionable crosstalk in long circuits. This crosstalk is guarded against by splicing successive lengths of cable in a random manner so that no two quads are adjacent for more than a small part of their total length. To reduce crosstalk between pairs in the same quad, the capacitive unbalance is measured and balancing capacitors are connected at some point in each loading section. In carrier systems applied to cable circuits, inductive coupling is more important as a cause of crosstalk. Indeed, the crosstalk possibilities at carrier frequencies are so great as to necessitate a number of basic changes in circuit design as well as the use of crosstalk balancing coils to reduce inductive crosstalk.

18.11 Applied mathematics.

A. Effect of Loading on Attenuation

The general attenuation constant of a transmission line is given by the expression:

α = Real part of $\sqrt{ZY}$, which equals

$$\sqrt{\tfrac{1}{2}\sqrt{(R^2+\omega^2L^2)(G^2+\omega^2C^2)}+\tfrac{1}{2}(GR-\omega^2LC)}$$

where α = attenuation constant per unit length
$Z = (R+j\omega L)$, the impedance per unit length
$Y = (G+j\omega C)$, the admittance per unit length
R = resistance per unit length
L = inductance per unit length
G = leakage conductance per unit length
C = capacitance per unit length.

With a non-loaded cable, satisfactory transmission is limited in practice to a few miles. In order to reduce the attenuation, the first solution which might present itself is the reduction of the resistance per unit length. However, this solution would prove uneconomical since it would result in the use of more copper per circuit and fewer circuits per cable. In well maintained lines, the leakage conductance G is kept at a minimum and can, for all practical purposes, be considered to be negligible. A reduction of the capacitance per unit length could only be secured by a wider separation of the wires; hence, this solution would also prove uneconomical since it would reduce the number of circuits which might be carried in a single cable and thus increase the cost per circuit. Therefore, to reduce the attenuation, there remains only the possibility of changing the value of the inductance per unit length, L. Neglecting leakage, the attenuation equation reduces to

$$\alpha = \sqrt{\tfrac{1}{2}\,\omega C\left[\sqrt{R^2+\omega^2L^2}-\omega L\right]}$$

A study of this expression shows that, within certain limits, an increase in L will result in a reduction of the attenuation per unit length. The practice of loading is thus merely a means of increasing the inductance per unit length of a circuit. The improvement that can be obtained by increasing L depends on the value of the resistance per unit length, R. If R is small, very little decrease in the attenuation can be effected by increasing L; if on the other hand R is relatively large, as it necessarily is in practical circuits, a substantial reduction in the attenuation can be effected. Inasmuch as any inductance that may be added has some resistance, loading will increase the resistance per unit length and to some degree neutralize the effort to improve conditions. But with proper design of the loading coils, the increase in L more than offsets the increase in R so that the attenuation constant is reduced. Theoretically, in order to increase L it would be necessary to increase the distributed inductance in the circuit since the equation for the attenuation constant is developed in transmission line theory on the basis of uniformly distributed line constants. This might be accomplished by wrapping each conductor with a spiral of magnetic tape but the cost would be so great as to render the method impractical.

In practice, the loading inductance is supplied in the form of coils inserted in the circuit at regularly spaced intervals. A circuit of distributed constants may be simulated by a series of T-networks of lumped constants. Thus the addition of inductances in lumped form will produce the effect of increasing the distributed inductance provided that the lumps are sufficiently close together.

B. *Effect of Loading on Distortion*

In non-loaded lines, neglecting the very small inductance and leakage between the wires, the impedance and the attenuation are given by the formulas

$$Z_o = \sqrt{\frac{R}{j\omega C}}$$

and,
$$\alpha = \sqrt{\frac{\omega RC}{2}}$$

From these two equations, it is seen that both the attenuation and the impedance vary with frequency giving rise to distortion effects. In a loaded cable, neglecting G, the reactance ωL has been made very large and the resistance R may now be considered insignificant as compared with the reactance. Thus, for the loaded cable, the equations for impedance and attenuation become

$$Z_o = \sqrt{\frac{L}{C}} \quad \text{(approximately)}$$

and,
$$\alpha = \sqrt{\frac{R^2C}{4L}} \quad \text{(approximately).}$$

It is seen from the above equations that both the impedance and the attenuation are substantially independent of frequency due to the introduction of a relatively large amount of inductance in the loading process. Hence, loading serves the dual purpose of reducing the energy loss and virtually eliminating distortion effects throughout the frequency range that is to be transmitted.

The loading coils in series with the line wires increase the line impedance. At the same time, they tend to neutralize the inherent line capacitance. Thus, the loaded line may be considered to be made up of a number of sections consisting of series inductance and capacitance, with each section acted upon by the voltages set up across the capacitance of the preceding section. This results in an increased voltage and a decreased current and practically eliminates the phase angle of the characteristic impedance of the line. Owing to the increased impedance, or the fact that the energy is transmitted at higher voltage and decreased current, the I^2R losses are less and thus the total energy loss must be less.

18.12 Questions and problems.

1. List the advantages and disadvantages of line communications as compared with radio communications.
2. How does the operational requirement influence the choice between telephone communication and telegraph communication?
3. Distinguish between a sound-powered telephone system and a common-battery telephone system.
4. Why are multiple switchboards required in large telephone exchanges?
5. Describe three methods of signalling used in telephone systems.
6. What two types of automatic dial systems are most widely used in telephone systems? Briefly distinguish between them.
7. Why are protective devices required in line communication systems? List the three principal types used.
8. What is the primary purpose of a plug-and-jack assembly?
9. Explain marking or spacing.
10. List the factors which limit the length of a land line.
11. Describe two principal types of regenerative repeaters used in teletype systems.
12. Define the following terms as used in line communications:
    (a) single operation
    (b) duplex operation
    (c) neutral operation
    (d) polar operation
    (e) polarential operation.
13. Define the following terms:
    (a) bias distortion
    (b) characteristic distortion
    (c) fortuitous distortion.
14. Describe how bias distortion may be eliminated in polar circuits.
15. Contrast the effects of bias and characteristic distortion in a polar telegraph circuit.
16. What is the purpose of loading line communication circuits and, briefly, how is this accomplished?
17. What secondary advantages result from loading land lines?
18. What are the disadvantages of loading?
19. List the factors that enter into the design of a loaded-line communication circuit.
20. Compare the relative merits of open-wire circuits and cable circuits and state the conditions which determine the use of each.
21. Outline briefly the elementary principles of carrier technique.
22. Distinguish between a low-frequency carrier system and a broad-band carrier system.
23. What methods are used to reduce noise and crosstalk on telephone lines?

CHAPTER XIX

TELETYPE

In discussions of telegraph operation by land lines or radio in previous chapters, it has been assumed that the electrical signals to be transmitted were produced by the manual operation of a telegraph key. A large percentage of telegraph circuits, however, are operated by mechanical devices known as *teletypewriters.* A teletypewriter link consists of two *teleprinter* units like the one shown in Fig. 19-1 supplied with suitable power and connected by a two-wire line. Fig. 19-2 shows the teleprinter unit with cover removed. It consists of a keyboard similar to a standard typewriter keyboard and printing mechanism which prints transmitted and received copy in page form.

Communication by teletype has a number of advantages over manual telegraph service. A transmission speed of 60 words per minute is generally used in teletype service on the North American continent, while speeds up to 100 words per minute are possible with modified equipment and circuits. In telegraph transmission by Morse code, a thorough knowledge of the code is required by the transmitting and receiving operators. Each letter or numeral being transmitted requires from one to five manual operations of the key. In teletype operation only one manual operation is required for each transmission character. Although a knowledge of the code is useful to the teletype operator, it is not essential.

19.1 Principles of teletype operation. Communication by teletype consists of reproducing by electrical means at a remote point the mechanical operations performed at the transmitting station. Each time the transmitting operator strikes a key to print a letter, the machine performs a series of mechanical operations and transmits a sequence of electrical impulses along the line to the receiver. These impulses operate a

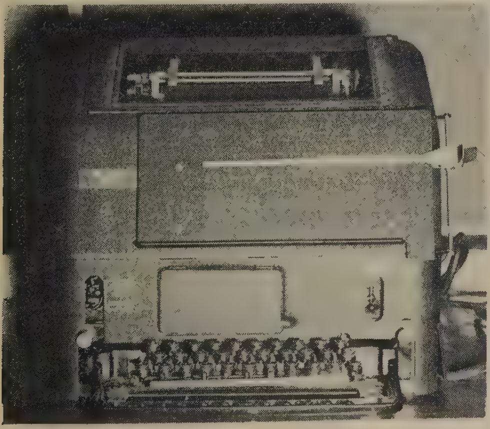


Fig. 19-1 Exterior view of teletypewriter unit.

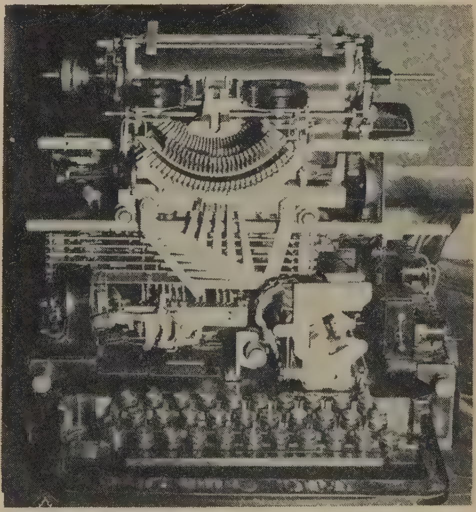


Fig. 19-2 Teletypewriter with cover removed.

Characters		Code signals						
L.C.	U.C.	Start	1	2	3	4	5	Stop
A	—	□	■	■	□	□	□	■
B	?	□	■	□	□	■	■	■
C	:	□	□	■	■	■	□	■
D	$	□	■	□	□	■	□	■
E	3	□	■	□	□	□	□	■
F	!	□	■	□	■	■	□	■
G	&	□	□	■	□	■	■	■
H	£	□	□	□	■	□	■	■
I	8	□	□	■	■	□	□	■
J	Bell	□	■	■	□	■	□	■
K	(	□	■	■	■	■	□	■
L	)	□	□	■	□	□	■	■
M	.	□	□	□	■	■	■	■
N	,	□	□	□	■	■	□	■
O	9	□	□	□	□	■	■	■
P	0	□	□	■	■	□	■	■
Q	1	□	■	■	■	□	■	■
R	4	□	□	■	□	■	□	■
S	'	□	■	□	■	□	□	■
T	5	□	□	□	□	□	■	■
U	7	□	■	■	■	□	□	■
V	;	□	□	■	■	■	■	■
W	2	□	■	■	□	□	■	■
X	1	□	■	□	■	■	■	■
Y	6	□	■	□	■	□	■	■
Z	"	□	■	□	□	□	■	■
Blank		□	□	□	□	□	□	□
Space		□	□	□	■	□	□	■
Car.ret.		□	□	□	□	■	□	■
Line feed		□	□	■	□	□	□	■
Figures		□	■	■	□	■	■	■
Letters		□	■	■	■	■	■	■

Signal lengths in msec. standard speed

40 speed	33	33	33	33	33	33	47
60 speed	22	22	22	22	22	22	31
75 speed	18	18	18	18	18	18	25

FIG. 19-3 Teletypewriter code.

mechanical selector system in the receiver which selects the corresponding type bar and causes the letter to be printed.

The code generally used in teletype systems is the *Baudot* code. It is a *five-unit* or *five-impulse* code and is illustrated in Fig. 19-3. It provides for transmission of the letters of the alphabet, the numerals, commonly used symbols and machine functions such as line feed and carriage return. Each code character consists of five units of equal duration. The black units of Fig. 19-3 represent *marking impulses* and the white units represent *spacing impulses.* These correspond to the circuit conditions of key closed and key open, respectively, in telegraph operation.

Teletypewriters are synchronized by means of a *start-stop* system. That is, the machines, instead of operating continuously are started at the beginning of each five-unit group and stopped at the end of the group. This ensures that the transmitting and receiving machines are always in exact synchronism at the beginning of the transmission of every character. The start-stop system so frequently corrects any timing differences that may exist that printing errors are not likely to occur. It requires, however, that two synchronizing current pulses be transmitted for each character. The first synchronizing pulse is a spacing impulse which precedes the five-unit group and sets the receiving machine in mechanical operation. The stop pulse is a marking impulse at the end of the five-unit group which stops the receiving mechanism. The start pulse is equal in duration to each of the five code impulses while the stop pulse is 1.42 times as long. The duration of each impulse corresponding to standard transmission speeds of 40, 60 and 75 words per minute is indicated at the bottom of Fig. 19-3.

The mechanical unit which transmits the series of electrical impulses when a key is struck is called the *distributor.* Several types of distributors have been designed and two are in common use in teletype equipment. They are the *commutator* or *segmented-ring distributor* and the *rotating cam distributor* described below.

A. Commutator or Segmented-Ring Distributor

This system, shown in diagrammatic form in Fig. 19-4, is representative of early teletype equipment and is more easily understood than later designs. The rotating brush arms of both the transmitting and receiving equipments are coupled to the driving shafts of two motors by friction clutches and are normally held stationary by the latches of the sending and receiving start magnets. The motors at the trans-

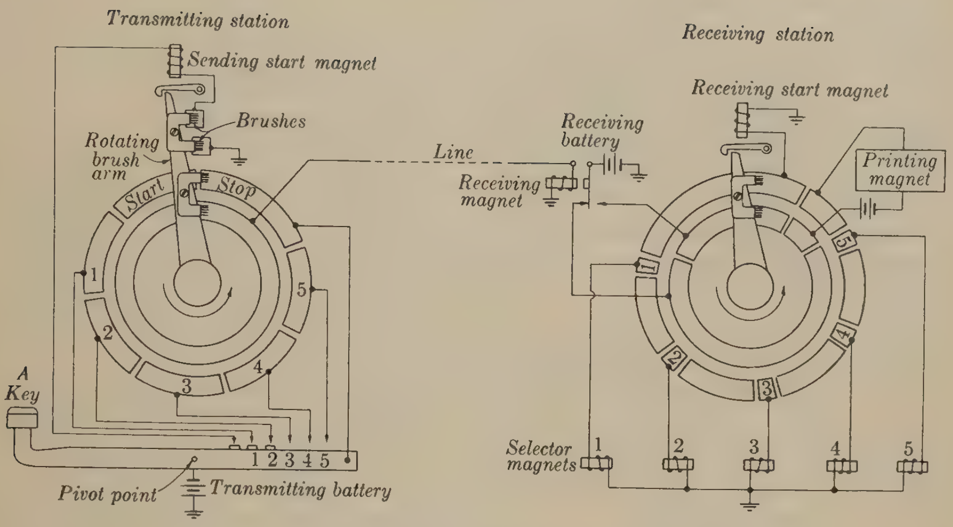


FIG. 19-4 Commutator or segmented-ring distributor.

mitting and receiving stations are governed to rotate at approximately the same speed. Each time a key is struck on the keyboard of the transmitting teletypewriter, the brush arm is released, makes one complete counterclockwise rotation and is brought to rest again in its original position. This causes the brush arm to be released in the teleprinter at the receiving station and to complete one revolution in synchronism with the transmitting brush arm. The electrical impulses which make up the letter code are sent over the line as the brushes rotate. Corresponding magnets are energized in the receiving unit to select the required printing bar and print the character. The operation is best understood by reference to a particular example. Suppose the letter *A* is to be transmitted. The electrical impulses corresponding to this character are start, mark, mark, space, space, space, stop.

1. START-STOP MECHANISM.

When key *A* is struck, the key lever rotates about the pivot point to close a number of electrical contacts as indicated in Fig. 19-4. There is then a path for electron flow from the transmitting battery, through the sending start magnet and the brushes on the brush arm to ground. The start magnet pulls up the latch and releases the brush arm. The brush arm is driven through one revolution by the motor until it reengages the latch. When the arm begins to rotate, the lower brushes move from the STOP segment to the START segment. The transmitting battery is then disconnected from the line and the magnet at the receiving end of the line is deenergized. The armature of this relay moves to the right and connects the receiving battery through the brushes and the receiving start magnet to ground. This latter magnet is energized, pulls up the latch and permits the receiving brush arm to rotate one revolution. When the transmitting brush arm comes to the end of its revolution, the brushes connect the transmitting battery through the STOP segment and the brushes to the line. The receiving magnet is energized and its armature moves to the left. This ensures that the receiving battery is disconnected from the receiving start magnet so the latch will engage the brush arm in the receiver and stop the motion at the end of one revolution. The receiving mechanism is then ready for a new sequence of operations. To ensure that the receiving brush arm will always complete its travel before a new sequence is initiated by the transmitter, the receiving brush arm is arranged to rotate about 14 per cent faster than the transmitting brush arm.

2. Code Impulse Mechanism.

When the teletypewriter key at the transmitting station is struck, contacts 1, 2, 3, 4 and 5 on the key lever are closed or left open according to the Baudot code for that letter. Thus, for the letter *A*, contacts 1 and 2 are closed while 3, 4 and 5 are left open. As the brushes rotate over segment 1 of the transmitting distributor, the transmitting battery is connected to the line. The receiving magnet is energized and its armature is moved to the left. When the brush arm of the receiving distributor sweeps over segment 1, the receiving battery is connected through the brushes to receiving selector magnet number 1. This magnet is energized and performs the first of the five mechanical operations required to select the correct type bar in the printing mechanism. Similarly, when the transmitting brush sweeps over segment 2, a marking impulse is again sent to the line. As the receiving brush contacts segment 2, the second selector magnet is energized. When the transmitting brush touches segments 3, 4 and 5, however, the transmitting battery is disconnected from the line and the armature of the receiving magnet moves to the right. The receiving battery is then disconnected from the inner commutator ring so that selector magnets 3, 4 and 5 are not energized when the receiving brush sweeps over the corresponding contacts. The selector magnets, energized according to the code, select the type bar carrying letter *A* in the teleprinter. Just before the receiving brush reaches the end of its travel, it passes over a pair of segments through which it completes the printing battery and magnet circuit to operate the type bar and print the letter.

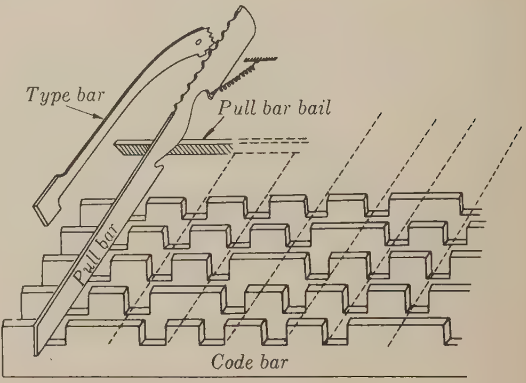


Fig.19-5 Selecting mechanism of a receiving teletype machine.

It is to be noted that the commutator segments are considerably smaller in the receiver than in the transmitter. This means that only a relatively small portion of each transmitted signal impulse is used for operation of the receiving machine. Such an arrangement permits some variation in the speed of rotation of the brush arms without introducing distortion and printing errors. As long as the receiver brush passes over a given segment while the transmitter brush is contacting the longer corresponding transmitter segment, normal operation will result.

B. Receiver Code Selector

Several types of code selectors for use in teletype receivers have been developed. Fig. 19-5 illustrates the operation of a code selector suitable for use with the segmented ring transmitting unit just described. The five selecting magnets referred to above control the motion of five *code bars*. If the first impulse following the start impulse is a marking signal, the first selecting magnet will

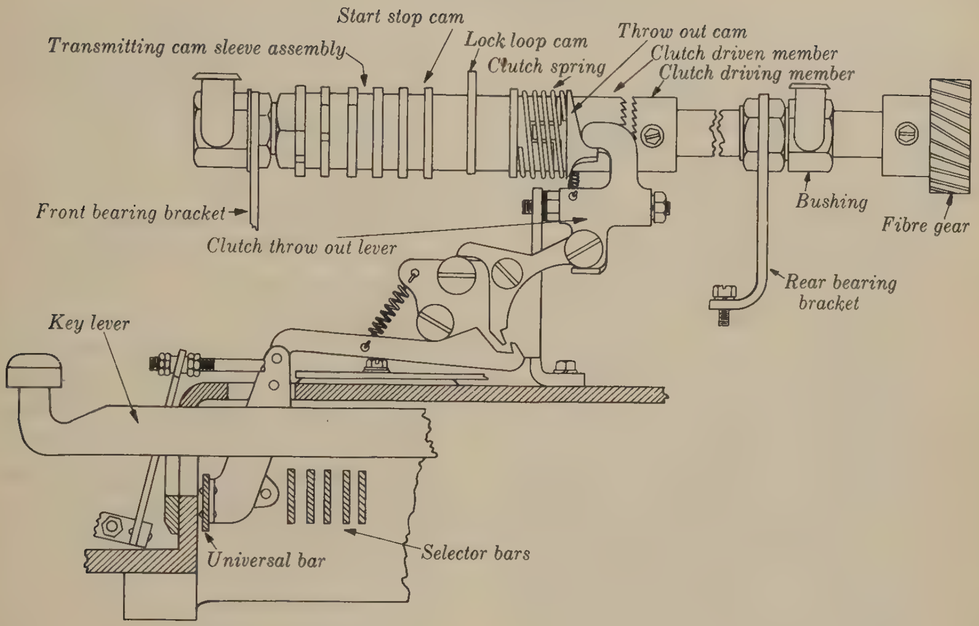


FIG. 19-6 Rotating cam transmitter assembly.

move code bar number 1 endwise a slight amount. Similarly code bar number 2 will be moved or left in position depending upon whether the second code impulse is a mark or a space signal and so on through the five impulses of the code. When all five impulses have been received, the code bars are so arranged that the slots under the particular *pull bar* corresponding to that particular code combination are in line, and all other groups are out of line. This one pull bar is then allowed to drop down a small distance where it rests on the *pull bar bail.* Then, when the printer magnet is energized, the pull bar bail moves forward and causes the *type bar* to print the character.

C. Rotating Cam Transmitter

While the commutator and rotating brush equipment is still widely used in certain units of the teletype system, other mechanical means have been devised for producing the electrical impulses which make up the five-unit code. The essential components of the transmitting mechanism are shown in Fig. 19-6. The components include the teletypewriter keyboard and a bank of electrical contacts which are actuated by rotating cams. The cams are mounted on a shaft which is driven at a constant speed by a synchronous motor. The cams may be held in a stationary position or may be allowed to rotate with the shaft by means of a friction clutch. When any key is depressed, the driven member of the clutch is permitted to move into mesh with the driving member, causing the cam sleeve assembly to revolve. At the end of the revolution, the driven member of the clutch is disengaged by the clutch and the cams are brought to rest until another key is depressed.

Mounted beneath the key levers are five *selector bars* and a *universal bar* extending across the width of the keyboard. The selector bars are provided with serrations as shown in Fig. 19-7, according to the requirements of the

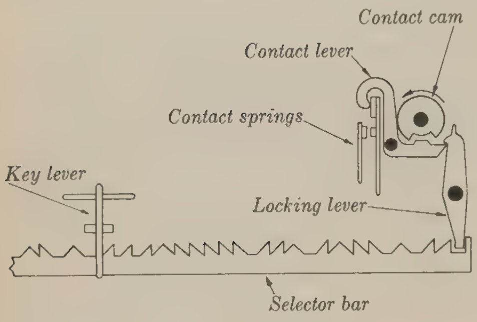


FIG. 19-7 Serrated selector bar and assembly.

signalling code. These bars rest on rollers and are guided at each end so that they may be easily moved endwise. When a key is depressed, the key lever strikes the slanting sides of these notches, moving the bars either to the right or left depending upon whether the impulses corresponding to the bars are to be marking or spacing impulses. If a spacing impulse is to be sent, the selector bar is held to the right as shown in Fig. 19-7. The *locking lever* engages the lower arm of the *contact lever* and keeps the contact springs apart as the cam rotates. When a marking impulse is transmitted, the key lever shifts the selecting bar to the left. The locking lever releases the contact lever so that it follows the indentation in the contact cam as the latter rotates. This permits the contact springs to move together to send a marking signal to the line.

The universal bar is mechanically coupled to the clutch. Thus, whenever a key or the space bar is depressed, the selector bars are set and the universal bar causes the cams to start rotating. Three sets of mechanical and electrical operations are performed by the rotating cams and are as follows.

(i) The *start-stop cam* opens a pair of contacts and sends a space impulse to the line. This impulse is the start element which starts a similar cam assembly rotating in the receiver. At the end of the revolution the start-stop cam closes the contacts and sends a marking impulse to the line. This impulse stops the rotation of the cam assembly in the receiver.

(ii) The *code-impulse cams*, five in number, operate the electrical contacts which provide the marking impulses of the five-unit code. The notches in these cams are so positioned as to actuate the contacts in succession as the cams rotate. The closing of the marking contacts is controlled not only by the cams but also by the positions of the five selector bars. Their operation is best understood by reference to a particular example. Suppose the letter *Y* is to be transmitted. The code for this letter is mark, space, mark, space, mark. Depressing the key bearing this letter causes the selector bars 1, 3 and 5 to shift toward one side of the keyboard and bars 2 and 4 to shift toward the opposite side. This shift causes contact levers 2 and 4 to be locked out of position. As the cam assembly revolves, cam number 1 actuates the corresponding lever arm and closes a pair of contacts. This operation sends a marking signal to the line. Cam number 2 is ineffective in actuating its lever arm because the latter was locked out of position by selector bar number 2. Thus a spacing signal is sent to the line during this interval. The remaining code-impulse cams operate in sequence, transmitting mark and space impulses according to the code.

(iii) In addition to the components described above, the transmitter contains a set of locking levers. These levers lock the selector bars in the positions determined by the key until the sequence of operations

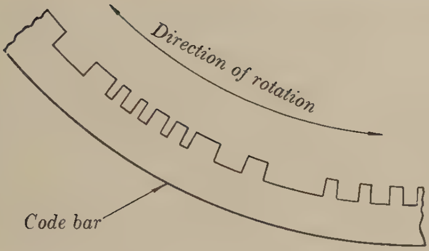


FIG. 19-8 Section of code bar in single magnet selector.

has been completed. At the end of the sequence the *lock-loop cam* unlocks the levers. The selector bars may then be re-positioned by depressing another key. The sequence of operations outlined above is summarized in tabular form on page 19:8.

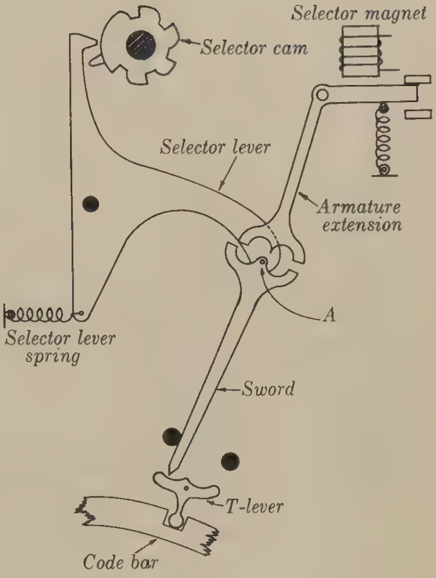


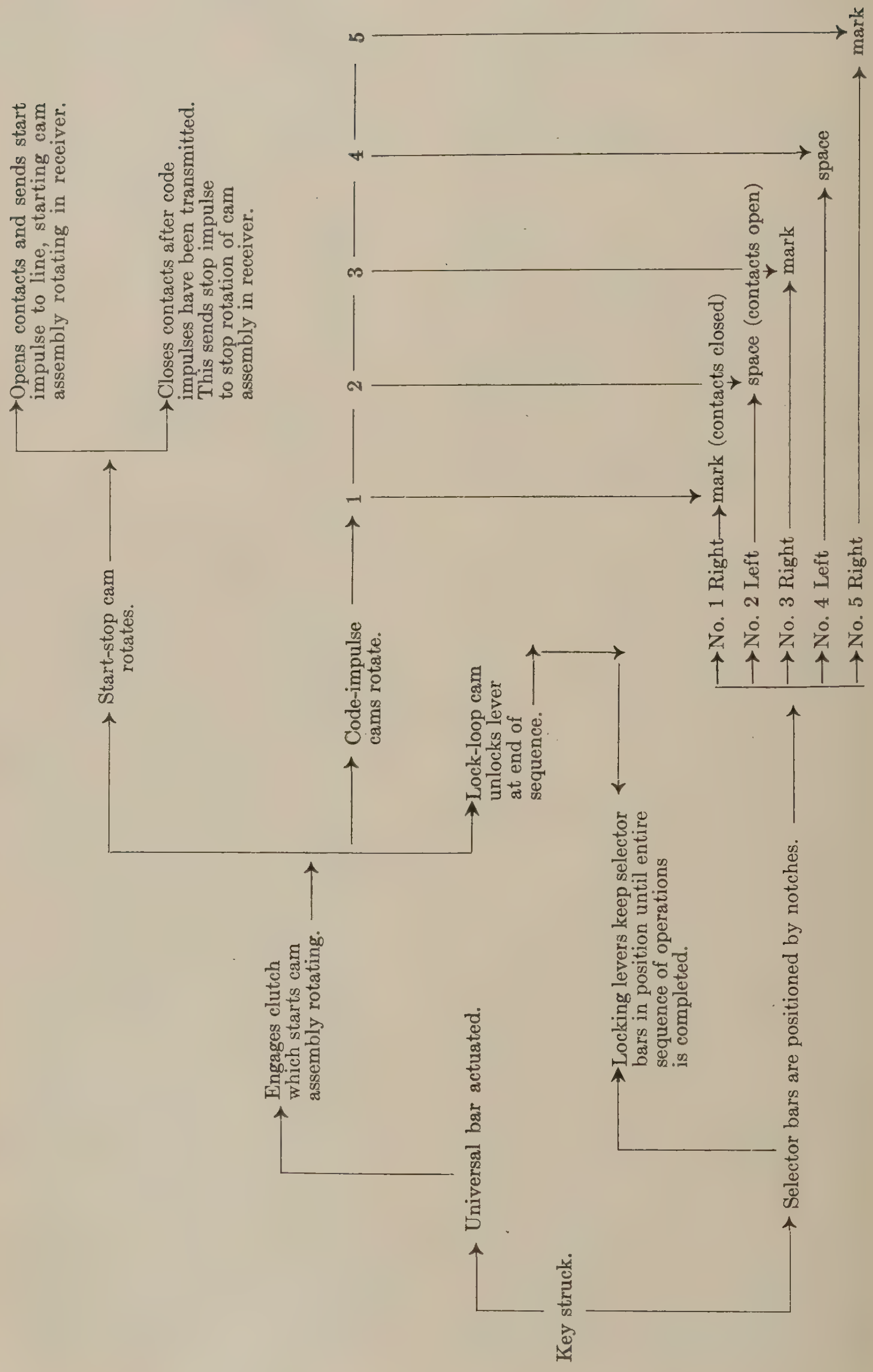
FIG. 19-9 Receiving mechanism of a single magnet selector.

D. Single Magnet Selector

Teletypewriters equipped with the rotating cam distributor in the transmitting mechanism also make use of a rotating cam assembly in the receiving unit. In this system the five electromagnets which operate the five code bars, are replaced by a single electromagnet. The code bars, instead of being straight and mounted to move endwise as shown in Fig. 19-5, are curved and mounted so that they can rotate as shown in Fig. 19-8. The essential mechanical and electrical components of the single magnet selector are shown in Fig. 19-9. The line signals from the transmitter are distributed by means of a group of six rotating cams so spaced angularly on a shaft that each will function at the time that the corresponding signal pulse is being received. Instead of closing electrical contacts, however, these rotating cams perform purely mechanical operations as follows. When the start pulse is received, the magnet armature is released. This operates a latch, not shown in the figure, and allows the shaft carrying the selector cams to start rotating. The cams are so spaced that at the time the first of the five impulses of the code signal is being received, the first cam engages the projection on the selector lever associated with the first code bar and rotates it slightly in a counterclockwise direction. The effect of this movement depends on whether or not the magnet armature is operated. If the first code impulse is a marking signal, the armature will be operated as shown in Fig. 19-9. The moving of the selector lever by the first selector cam lifts up the *sword* and causes the right hand projection on its upper end to strike the right hand end of the *armature extension*. This rotates the sword in a clockwise direction about its pivot *A*. When the selector cam in its continued rotation clears the *selector lever* and allows the selector lever spring to restore it to its normal position, the point of the sword is brought down against the left hand side of the *T-lever*, rotating it in a counterclockwise direction and so moving the *code bar* to the right.

If the line signal from the transmitter

Sequence of Operations in Transmitting Letter Y by Teletype.


Key struck.
Universal bar actuated.
Engages clutch which starts cam assembly rotating.
Start-stop cam rotates.
Opens contacts and sends start impulse to line, starting cam assembly rotating in receiver.
Closes contacts after code impulses have been transmitted. This sends stop impulse to stop rotation of cam assembly in receiver.
Code-impulse cams rotate.
1
2
3
4
5
Lock-loop cam unlocks lever at end of sequence.
Locking levers keep selector bars in position until entire sequence of operations is completed.
Selector bars are positioned by notches.
No. 1 Right
No. 2 Left
No. 3 Right
No. 4 Left
No. 5 Right
mark (contacts closed)
space (contacts open)
mark
space
mark

is a spacing signal, the magnet armature is not actuated. Then, when the selector lever raises the sword, its left hand projection strikes the left hand side of the armature extension, causing the sword point to move to the right and the code bar to the left. In exactly the same manner, when the second code impulse is received, the second selector cam will have arrived at the proper position in its rotation to operate the selector lever associated with the second code bar, and it will be positioned according to the position of the magnet armature at the time. After all five signal pulses have been received and the code bars properly positioned, the sixth cam releases a clutch and allows the printing mechanism to operate.

While the foregoing discussion refers to the transmitting and receiving mechanisms as separate units, they are in fact both included in the same teletypewriter unit as shown by the photographs of Fig. 19-1 and Fig. 19-2. Such an arrangement makes it possible to connect machines for simplex, duplex or half-duplex operation. As in the case of telegraph practice, simplex operation refers to sending teletype signals over the same conductors and at the same time as telephone signals are being transmitted. At the receiving end of the line the teletype and telephone signals are separated and are channelled to their respective receiving units. With a simplex system it is possible to transmit in only one direction at one time. The receiving operator must wait until the sending operator has completed his transmission before he may start transmitting. A duplex system provides independent transmission on two separate paths so that traffic may be passed in both directions simultaneously. In a half-duplex system, the terminal equipment is capable of duplex operation but the line facilities permit transmission in only one direction at one time. Thus a half-duplex system will handle the same amount of traffic as a simplex system. or half the traffic of a full duplex system.

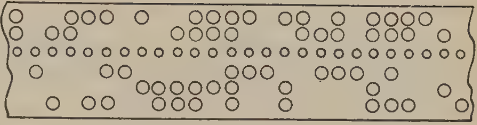


FIG.19-10 Perforated teletypewriter tape.

19.2 Perforated tapes. In order to maintain the teletype system in operation at its maximum efficiency, it is convenient to prepare traffic in advance for transmission at a routine time. This is accomplished by recording messages in the form of coded perforations on a paper tape. The tape may be read and checked for accuracy before being fed into a *transmitter-distributor* unit for automatic transmission by land lines or radio.

A section of a perforated teletypewriter tape is shown in Fig. 19-10. The small holes are feed holes which are used to feed the tape through the perforator and transmitting device. The tape may be read by moving it from right to left with the rough side up and with two recorded impulses above the feed line and three recorded impulses below the feed line. The tape is perforated for marking signals and is left unperforated for spacing signals. Start and stop signals are not recorded on the tape.

A. Perforators

The perforating mechanism consists essentially of a typing keyboard, a selecting mechanism, a punch hammer and a set of punches for perforating the tape. The punch hammer is controlled by a pair of electromagnets. When the key is struck to perforate the letter *A* (mark, mark, space, space, space) punch bars 1 and 2 are left in the path of their respective punches while punch bars 3, 4 and 5 are moved away from their punches. Then, when the punch magnet is energized, the hammer causes punches 1 and 2 to perforate the tape while punches 3, 4 and 5 are not operated. A feed hole is also punched with each forward motion of the punch hammer.

B. Reperforators

A *reperforator* is a motor-driven machine which receives electrically transmitted signals and *converts* these signals by mechanical means into code combinations perforated in a paper tape. In addition the message being received is printed on the tape. It is important to note that the perforations representing the printed letter are always six positions in advance of the letter. The reperforator is used to convert incoming messages, which would normally have to be retransmitted manually, into tapes which can be fed directly into the transmitter-distributor. This cuts down the handling of traffic and increases the speed with which messages can be relayed.

Another useful device in the teletype centre is the *typing reperforator*. This machine is fitted with a keyboard for preparing tapes for subsequent transmission. In addition it may be used for perforating tapes from signals received from a distant station. It thus functions as either a perforator or as a reperforator.

C. Tapes

Two different types of tape are used in teletype practice. The perforator in punching the tape leaves a circular hole by completely removing the tape as shown in Fig. 19-10. The pieces of tape removed by this operation are called *chads* and the tape is known as *chad tape*. The reperforator or typing reperforator does not completely remove the circular piece of paper but leaves it in the tape secured by a small uncut portion. This type is called *chadless tape* and is shown in Fig. 19-11.

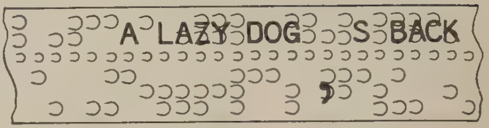


Fig. 19-11 Partially perforated (chadless) teletypewriter tape.

D. Transmitter-Distributor

The perforated tapes are fed into the *transmitter-distributor* unit for transmission over land line or radio links. When using the transmitter-distributor, no printed copy is produced at the transmitting end unless a separate printer is put into the line for this purpose. The tape is fed under a clip and over a line of five *sensing pins*. These pins open or close electrical contacts according to the perforations in the tape. Each pair of contacts is connected in series with one segment of a commutator-type distributor similar to that shown in Fig. 19-4. A motor-driven revolving brush sweeps over the commutator segments to open and close an electrical circuit according to the code for the character being sensed by the pins. The brushes connect the transmitting power to the line during marking signals and disconnect it during spacing signals.

E. Automatic Numbering

In centres where traffic is received on tape for relaying to other stations, it is important that a definite procedure be used for identifying messages and for giving a constant check on possible omissions and errors. Normally, messages handled direct from station to station bear a serial number making it possible for both stations to check on the number of messages received and transmitted. When a message is relayed manually, it receives a new serial number.

In tape relay centres, an automatic numbering device is used for inserting a *message identifying number* and a *channel*

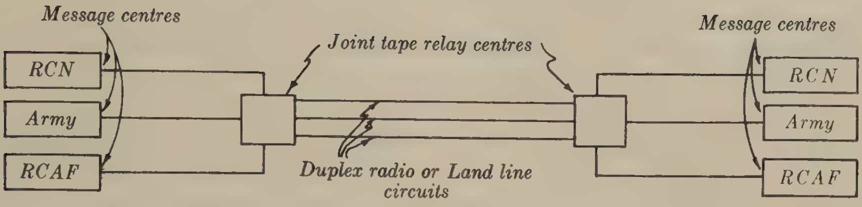


FIG. 19-12 Service joint tape relay system.

letter on the tape for each message passed. Messages originating at a given centre are automatically numbered consecutively. If two or more channels exist between the receiving and transmitting centres, a channel letter is added. Where messages are relayed the message identifying number is retained but preceded by a channel message number with a channel letter if necessary. At specified intervals, normally every twenty-four hours, each station to which traffic has been passed checks on the messages received during the preceding interval. A new sequence of message identifying numbers is commenced at the same time.

19.3 Tape relay systems. A highly efficient system of passing messages using teletype techniques has been developed in the form of the *semi-automatic tape relay system*. A semi-automatic tape relay system consists basically of two or more message centres which relay or regenerate teleprinter messages through one or more tape relay rooms to other message centres. This system is widely used both by commercial companies and by the Services. In Canada, the National Defence Communication System utilizes *joint tape relay centres* as illustrated in Fig. 19-12.

Each message centre is responsible for transmitting messages to the tape relay room by means of tape or keyboard-operated teleprinters employing duplex land-line teletype circuits. Messages are received in the tape relay rooms on tape only and in message centres on page printing machines. Tape relay rooms are connected by a number of teletype circuits, either radio or land line. The tape relay system has many advantages which are as follows.

(i) Economies of space and building costs are effected by having long-line terminal equipment under one roof.

(ii) Larger traffic capacities and faster handling of messages are possible.

(iii) Under conditions of heavy traffic, alternate routes are available for messages.

(iv) The human error factor in transcribing messages at relay points is eliminated; the messages are received at their destination in exactly the same form as they were teletyped by the originator.

(v) A comparatively few trained operators can handle a large volume of traffic.

The layout of a Service tape relay centre is shown in Fig. 19-13. *Receiving cabinets*, installed in the centre, contain typing reperforators along with their associated relays and power supplies. The individual machines are connected to incoming lines and record the incoming messages on chadless tape. Each *transmitting cabinet* contains two 3-gang transmitter-distributors which may be operated separately or in tandem. This allows up to six messages to be automatically transmitted on a circuit in quick succession. The *number transmitting cabinets* each contain one 3-gang transmitter-distributor having long, durable, perforated tapes punched with

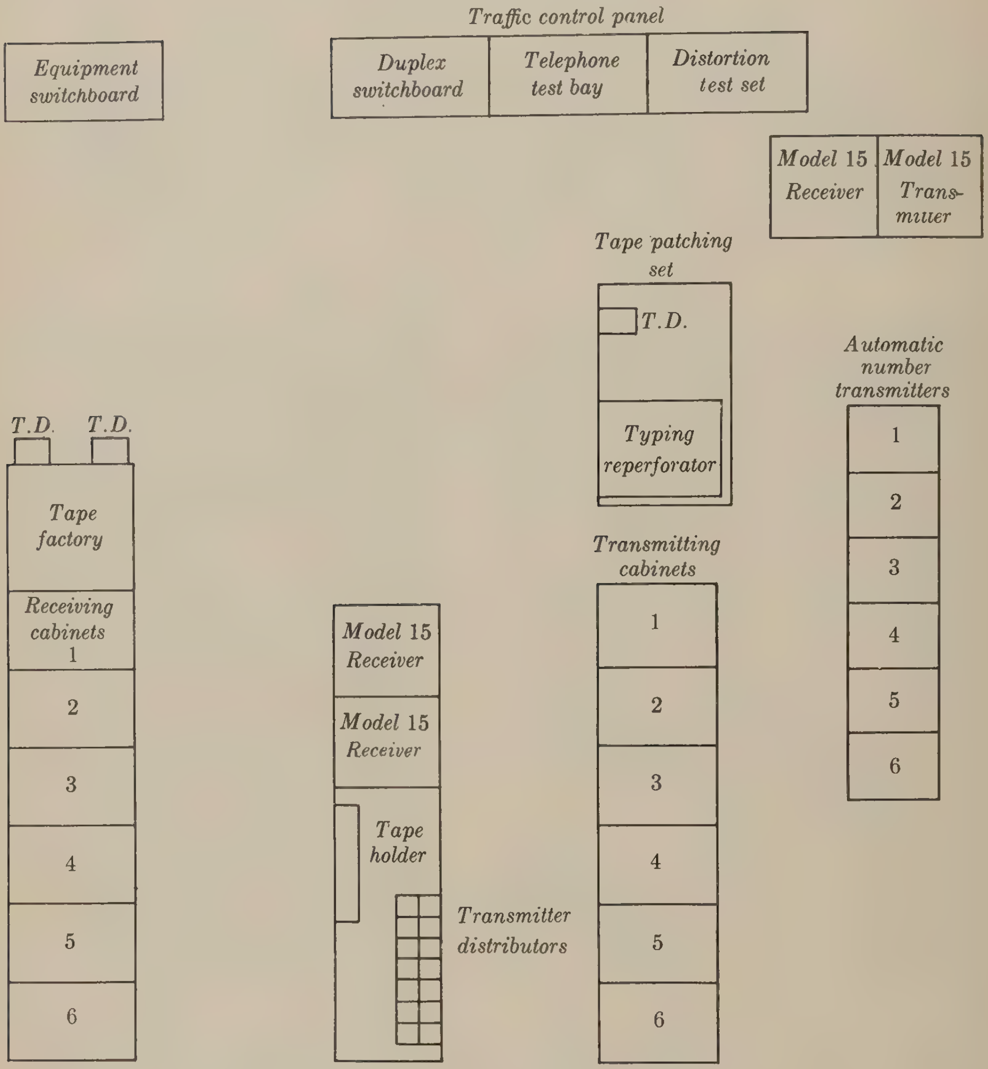


FIG.19-13 Layout of Service tape relay centre.

channel designations and successive numbers. When a message is placed in a transmitter it automatically starts the number transmitter associated with it. This unit, in turn, inserts the correct channel designation and serial number on the outgoing message. The number transmitter then stops and the message tape recommences feeding into the circuit. In a similar manner, both the time and date may be added to every outgoing message. Associated with each number transmitting cabinet is a power supply and a bank of relays. *Monitor cabinets* contain typing reperforators with associated power supplies to make a tape copy of every outgoing message for file purposes.

When a message which requires distribution to several addressees is received, the received tape is fed into a transmitter-distributor in the *multiple address cabinet,* the output of which may be connected or *patched* into several typing reperforators. This permits simultaneous manufacture of several tapes which can be used to send the message to its various destinations simultaneously and with a minimum of delay. The switchboard monitor set is located adjacent to the equipment switchboard and contains a typing reperforator with a keyboard and a single transmitter-distributor with necessary power supplies and plugs and keys for patching into the switchboard. When tapes are multilated or indistinct, they must be repaired before re-transmission. The defective tape is fed into the transmitter-distributor which operates the typing reperforator; when the mutilated section is reached, the transmitter-distributor is stopped and the section re-typed on the keyboard of the reperforator. When this has been completed, the tape in the transmitter-distributor is moved past the defective portion and the transmitter-distributor started again to complete the manufacture of the new tape. Model 15 teleprinters are located in the traffic control room near the traffic control panel. These machines are used for communicating with message centres and other stations where difficulties are encountered or orders must be given in connection with traffic control.

19.4 Operating speeds and bandwidth requirements. Three methods of indicating the speed of transmission of teletype signals are in common use.

1. WORDS PER MINUTE.

In indicating speeds of transmission, a standard word is considered to consist of six characters. Spacing between words and machine functions such as carriage return represent additional operations that must be performed by the operator. The speed with which traffic is passed by manual operation of teletypewriters depends upon the skill of the operator until the limiting speed of the machine is reached. In automatic transmission using perforated tapes, an operating speed of 60 words per minute is commonly employed.

2. CYCLES PER SECOND.

As indicated in Fig. 19-3, a teletype operating speed of 60 words per minute requires the transmission of five code impulses, each of 22 *m*sec. duration together with a start impulse of 22 *m*sec. and a stop impulse of 31 *m*sec. duration. Each signal element may be considered to be one half cycle of a square waveform alternating current. The duration of a complete cycle is therefore 44 *m*sec. and the number of cycles per second is $\frac{1000}{44}$ or approximately 23. The square waveform is built up of a sinusoidal fundamental of 23 c/s plus odd harmonics of reduced amplitude. Teletype equipment operates satisfactorily if the fundamental and third harmonic (69 c/s) are transmitted. This requirement must be met in carrier teletype systems employing land lines and radio links.

3. BAUDS.

Teletype transmitting speeds are sometimes stated in *bauds*. The speed in bauds equals the number of signal elements of shortest duration transmitted per second. Thus a speed of 60 words per minute or one signal element every 22 *m*sec. is equal to $\frac{1000}{22}$ or 45 bauds. The speed in cycles per second and in bauds corresponding to any other number of words per minute may be calculated in a similar manner.

19.5 Adaptation of radio to teletype circuits. Several methods of adapting standard radio communication transmitters and receivers to teletype systems have been developed. Such a system, known as *radio teletype*, not only eliminates long land lines, but also makes the installations mobile. As a result, the high-speed operation and direct-printing facilities inherent in teletype are available for ships and aircraft.

Before a satisfactory service can be achieved, it is desirable that these radio links be made as reliable as land lines and be designed so that they can be incorporated in the networks without recourse to manual transcription or techniques which will complicate message handling.

In adapting radio to teletype operation, the current pulses comprising the teletype code must be converted to some form of modulation of an r-f carrier. Several methods of modulating the r-f carrier with teletype signals have been developed.

A. On-Off Keying

Since there are only two conditions required by the machines, i.e. current flow (mark) and no current flow (space), the logical procedure would be to have the transmitter silenced during the idle periods of the machine and radiating energy during the operating periods. Also, since the machine normally rests in the mark position, this condition would correspond to no radiation from the transmitter. This system, however, presents some difficulty in providing automatic gain control should this facility be desirable.

B. Reversed On-Off Keying

As the name implies, this type of operation is the reverse of on-off keying. The radio transmitter radiates energy during the idle periods of the teletype machine and intelligence is conveyed by interrupting this transmission when the machine operates. Such a system ensures that a signal is present in the receiver to prevent the operation of the automatic gain control on noise signals alone during the relatively long idle periods of the machine.

C. Frequency-Shift Keying

The type of modulation finding widest application in the Service where one channel is required is *frequency-shift keying* (F.S.K.). It is arranged that the mark and space elements are transmitted on two discrete frequencies. The frequency separation at present is 850 c/s on the high-frequency communications band and 200 c/s on the low-frequency communications band. With frequency-shift keying there is no amplitude modulation at the transmitter, but since both mark and space signals are transmitted there is a form of frequency modulation of fixed frequency deviation. Regardless of the type of keying employed, each receiver must be followed by a stage which will convert the output of the radio receiver into current pulses to operate the teletypewriter associated with the equipment.

Since frequency-shift keying is generally employed in Service applications, the discussions which follow will refer to this type of radio teletype unless otherwise noted.

19.6 Principles of radio teletype systems. A block diagram of a radio teletype link is shown in Fig. 19-14. The transmitting portion consists of a teletypewriter, a frequency-shift keyer and a communications transmitter supplied with suitable power. The output signal from the teletypewriter consists of mark and space impulses according to the Baudot code described in section 19.1. The line current is normally $60mA$ for mark and $0mA$ for space impulses in neutral systems. In polar systems, the current is $30mA$ in one direction for mark and $30mA$ in the opposite direction for space signals.

The frequency-shift keyer is used in place of the oscillator stage in the transmitter. It produces an r-f carrier of constant amplitude which provides the drive to the transmitter. The frequency of the carrier is shifted 850 c/s from the mark to the space condition for operation in the high-frequency communication band. Thus, if the assigned or *phantom* carrier frequency is 10 Mc/s, the mark frequency is 10,000,425 c/s and

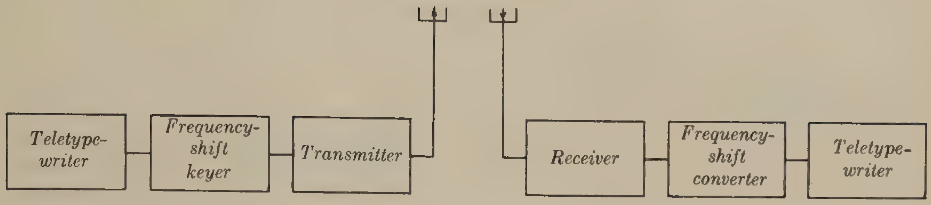


FIG.19-14 Block diagram of radio teletype link.

the space frequency 9,999,575 c/s. For operation in the low-frequency communication band, the frequency shift is 200 c/s. Since the difference between the mark and space frequencies is always less than 1000 c/s, it is essential to use crystal control of the signal frequency. It is customary that when the system is idle, it rest in the mark condition; that is, the output frequency is higher than the phantom frequency by half the total frequency shift when no intelligence is being transmitted.

The frequency-modulated output of the frequency-shift keyer is fed into the driver and power amplifier stages of a standard communications transmitter for transmission to distant receivers. It is important to note that if frequency multiplication occurs in the transmitter, the amount of frequency shift applied at the keyer will be multiplied in the transmitter. Thus, if frequency doubling occurs in the transmitter, the amount of shift applied in the keyer must be half the required final shift in the transmitter output.

The receiving equipment of the radio teletype link consists of a standard communications receiver, a frequency-shift converter and a teleprinter, together with power supplies. To ensure reliable operation of the receiver, a high-stability crystal-controlled local oscillator is used. The frequency of the b.f.o. is adjusted so that the output from the second detector is in the audio range or slightly higher. This signal shifts in frequency for mark and space impulses by the same amount as that applied at the transmitter. The signal is then passed to the *frequency-shift converter* which discriminates between the mark and space frequencies and converts them to current pulses which drive the teletypewriter.

For single or half-duplex operation of the radio teletype link, the terminal equipment at each end of the system includes a single-unit teletypewriter with page printer and a common aerial for the transmitter and receiver. A send-receive switch is used to connect the teletypewriter and aerial to the transmitting and receiving network as shown in Fig. 19-15.

From the foregoing brief survey, it is apparent that the frequency-shift keyer and the frequency-shift converter are the two units peculiar to teletype com-

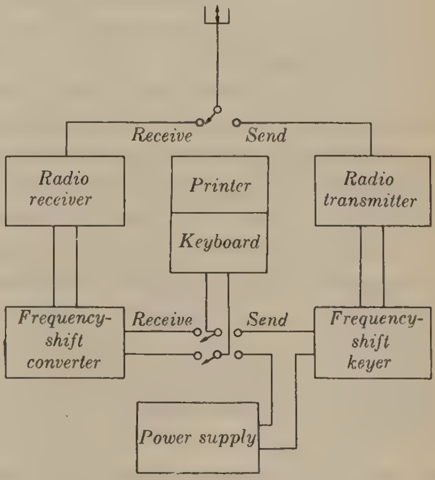


FIG.19-15 Block diagram of terminal equipment for radio teletype link.

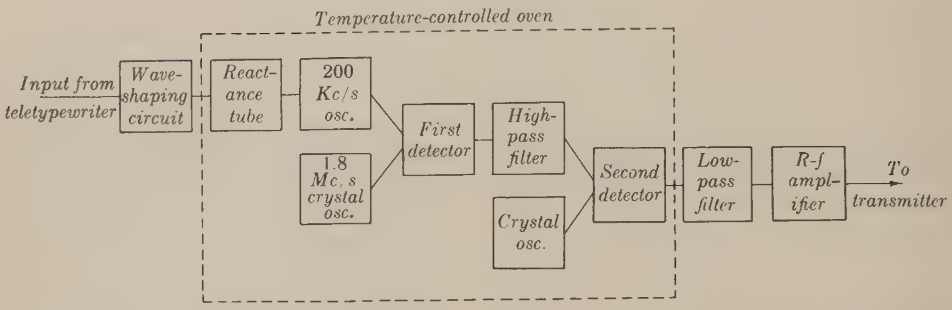


Fig. 19-16 Block diagram of low-frequency keyer circuit.

munication using radio links. The following sections are devoted to a more detailed description of these units of the radio teletype system.

19.7 Frequency-shift keyers.

A. Low-Frequency Keyer

A frequency-shift keyer for use in the low-frequency band is shown in block diagram form in Fig. 19-16. It is used to drive communications transmitters operating in the range 50 Kc/s to 500 Kc/s. The transmitter frequency is shifted from a lower space frequency to a higher mark frequency by the output of the teletypewriter. The amount of frequency shift is 200 c/s.

The required transmitter frequency is derived by the heterodyning of higher frequencies. The square-wave signal from the teletypewriter is passed through a wave-shaping circuit to a reactance tube. The wave-shaping circuit limits the amplitude of the keying voltage to make it independent of variations over a considerable range. It also impresses the proper polarity of keying voltage on the reactance tube. The reactance tube circuit is used to shift the frequency of a 200 Kc/s Colpitts oscillator 100 c/s above and below its normal frequency for mark and shift conditions. The output of the Colpitts oscillator is then combined with the output of a crystal oscillator operating at 1.8 Mc/s in a balanced modulator. Detection takes place and the upper sidebands are selected by high-pass filters to obtain a carrier at 2 Mc/s with a frequency shift of plus and minus 100 c/s for modulation. This signal is again combined with the output of a second crystal oscillator which operates at a fixed frequency in the range 2050 Kc/s to 2500 Kc/s. A selector switch makes it possible to select any one of three crystals in this range. The second detector is followed by low-pass filters which select the difference frequency to give a signal in the range 50 Kc/s to 500 Kc/s with a frequency shift of 200 c/s.

The r-f carrier is amplified by untuned amplifiers which pass the signal to the transmitter. Any harmonic distortion in the r-f carrier is largely eliminated by the Class *C* amplifiers in the transmitter. The frequency-controlling circuits of the keyer are contained in a temperature-controlled oven to maintain the frequency of the carrier constant.

B. High-Frequency Keyer

The diagram of a high-frequency keyer is shown in Fig. 19-17. It produces a carrier signal in the range from 2 to 5.5 Mc/s with a frequency shift up to 1000 c/s and may be used for the transmission of teletype, telegraph, f-m telephone, facsimile or telephoto communication.

1. Keying and Wave-Shaping Circuit. Depending on the particular installation, the input from the teletypewriter

keyboard may take one of four distinct forms. The two basic forms result from polar and neutral operation. In addition the input may vary as to the direction of current flow for a given mark or space condition. The input circuit of the keyer includes a sense switch S_6 to permit operation from either polar or neutral signals and a limiter to compensate for the differences in voltage amplitude resulting from the two types of keying. Also included in the keyer stage are two switches, S_4 and S_5. S_4 is a three-position switch used to simulate either the mark or space condition for testing. This switch is normally left in the line position so that the input from the keyboard controls the keyer. The FAX—F.S. switch S_5, in the FAX position provides for photo-facsimile operation of the keyer. It is set in the F.S. position for frequency-shift operation.

2. CRYSTAL OSCILLATOR.

The basic frequency of the keyer is derived from a Pierce crystal-controlled oscillator V_2. Provision is made to select any one of three output frequencies by a crystal selector switch S_3.

3. 200 KC/S OSCILLATOR.

To provide facility to shift the output frequency, the output of a stable 200 Kc/s oscillator V_8 is combined with that of the crystal oscillator. The 200 Kc/s oscillator is a balanced Colpitts oscillator operated at a relatively low level for stability.

4. REACTANCE TUBE.

V_5 is connected across C_{19} and C_{19A}, one half of the oscillator tank circuit in an inductive reactance-tube modulator circuit. A voltage from the other half of the oscillator tank is applied to V_5 grid through the phase-shifting network C_{33} and R_{24} in series. C_{25}, a .01μF capacitor offers only a negligible reactance and the lower end of R_{24} can be considered to be at earth potential for r-f. Thus, applied to the grid of V_5 is a voltage lagging that across C_{19} and C_{19A} by an angle slightly greater than 90°. Current through V_5 resulting from the grid voltage and the voltage applied to its anode via C_{24} thus lags the voltage across C_{19} and C_{19A} by 90°. V_5, when conducting, thus appears as an inductance connected in parallel with C_{19} and C_{19A}. When V_5 is not conducting, this apparent inductance is removed from the circuit reducing the oscillator frequency. In teletype operation the reactance tube is operated from cut-off to conduction for space and mark respectively; the effect is the same as if a small fixed inductance were switched in or out of the oscillator tank circuit. Capacitor C_{32}, between V_5 cathode and ground, forms the wave-shaping circuit, rounding off the keying pulses to restrict sideband radiation.

In the F.S. position of S_5 the cathode of the reactance tube is connected through R_{46} and R_{25} to the junction of R_{26} and R_{23} which form a voltage divider across the regulated 105-volt supply. Ignoring any effect of the keyer valve, anode current will flow through V_5, a steady operating bias being developed across R_{25} and R_{46}. These are the conditions which prevail when the left hand side of V_9 is non-conducting, conduction of V_5 resulting in an increase in output frequency for the mark condition. With the left side of V_9 conducting, the increase in current through R_{25} and R_{46} raises V_5 cathode sufficiently above grid potential that its current is cut off. V_5 is then no longer active in the 200 Kc/s oscillator circuit resulting in a decrease in frequency. This is the normal space condition. Thus, to control the reactance-tube modulator V_5, the left side of V_9 is keyed on and off. This is accomplished by the right side of V_9 and its anode load R_{30}. The cathode of the left side of V_9 is positive by the voltage across R_{46}, R_{25} and R_{23} when conducting and positive by the

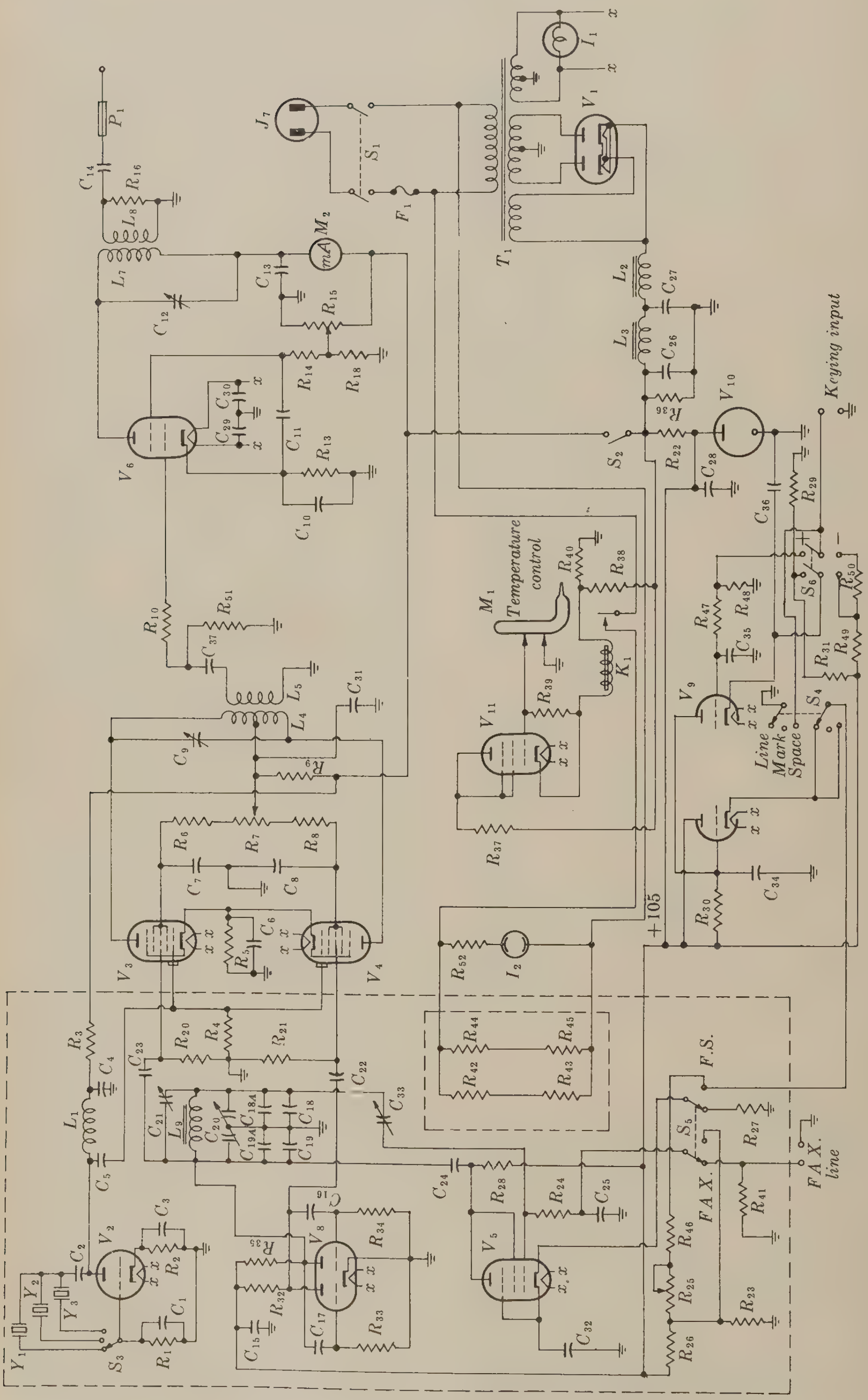


FIG. 19-17 Frequency-shift keyer for high-frequency operation.

voltage across R_{23} when non-conducting; its anode is connected to the regulated 105-volt supply; its grid is at the anode potential of the right side of V_9. Assume S_6 to be in the upward position which is for the reception of pulses of current, the direction of which produces a positive voltage at the top of R_{48}. It is immaterial whether these pulses are polar or neutral; either type is of sufficient amplitude that the right side of V_9 is driven to saturation by the positive voltage on its grid resulting in limiting action. Anode current through R_{30} drives the left grid of V_9 below cut-off placing the keyer in the mark condition. In the space condition, the right grid of V_9 is either at zero potential for neutral or at a negative potential for polar. In either case, the grid is sufficiently negative to the cathode to cut the tube off. The cathode is held positive by the voltage across R_{29} which is in series with R_{31} forming a bleeder across the 105-volt regulated supply. No anode current flows through R_{30} and the left side of V_9 will conduct, cutting off the reactance tube V_5 resulting in the space condition.

Should the teletype installation be such that the direction of current flow is opposite to that described above, S_6 is placed in the opposite position. Now the cathode of the right side of V_9 is connected to the 105-volt regulated supply through R_{49} and to the input line from the keyboard through R_{50}. With the reception of a marking impulse, electron flow is through R_{50} and R_{49}. The negative voltage at the junction of R_{49} and R_{50} places the cathode negative to the grid allowing current to flow. This drop in potential for either polar or neutral keying is sufficient to produce saturation limiting. In the space condition, the right side of V_9 is cut off. The action of the remainder of the circuit is unchanged. If it should be required to change the sense of the transmitted signal, that is, have the low-frequency condition correspond to the mark and the high-frequency correspond to the space, S_6 can be placed in the opposite position.

5. Balanced Modulator.

Tubes V_3 and V_4 are connected in a balanced modulator circuit. Applied to the control grids in parallel is the output from the high-frequency crystal oscillator. The output of the 200 Kc/s oscillator V_8 is applied to the signal grids in push-pull. The crystal frequency is cancelled out because the components of anode current at this frequency flow through the tank coil L_4 simultaneously in opposite directions producing cancellation in the transformer secondary. The tank circuit, L_4 and C_9, is tuned to the sum of the two frequencies. It therefore offers only a low impedance to the difference frequency which is 400 Kc/s below the desired frequency. Thus applied to the amplifier V_6 is the sum of the frequencies of the two oscillators. The output of V_6 is of sufficient amplitude to drive the associated transmitter.

6. Oven Temperature Control.

The three frequency-determining circuits, high-frequency oscillator, 200 Kc/s oscillator and reactance-tube modulator are mounted in a temperature-controlled oven. The temperature of the oven is controlled by a mercury thermometer M_1 with two contacts inserted into the channel. Relay K_1 and the control valve V_{11} complete the temperature-control circuit. The cathode of V_{11}, connected to the junction of R_{40} and R_{38}, a voltage divider across the 105-volt regulated supply, is at a positive potential. The grid is connected to the cathode via R_{39}, and to one of the contacts of the thermometer, the other contact of the thermometer being connected to earth. Relay K_1 in the V_{11} cathode circuit is energized when anode current flows, and de-energized when the valve is cut off.

The contacts of K_1 are in series with the mains supply to the heater elements R_{42}, R_{43}, R_{44} and R_{45}. If the temperature in the oven is below 60° C., the mercury in the thermometer contracts, opening the circuit between the contacts in the mercury channel; V_{11} grid is then at cathode potential allowing the tube to conduct, energizing relay K_1 and completing the supply to the heating elements. When the temperature exceeds 60° C., the mercury in the thermometer expands, completing the circuit between the two contacts through the mercury. K_1 is then de-energized breaking the supply to the heater elements. Relatively fast cycling occurs and the actual temperature variation inside the oven is of the order of a few hundredths of a degree centigrade.

7. POWER SUPPLY.

The power supply is entirely conventional.

8. FUNCTION OF CONTROLS.

The various controls and their functions are as follows.

(i) The MIXER TUNING control operates C_9, the capacitor in the tank circuit of the balanced modulator stage. The tank circuit is tuned to the sum of the two oscillator frequencies.

(ii) With the FREQUENCY control C_{20}, the frequency of the 200 Kc/s oscillator can be varied a few Kc/s. Since the output frequency is directly related to the 200 Kc/s oscillator frequency this control varies the output frequency for both the mark and space conditions.

(iii) To vary the magnitude of the frequency shift, R_{25}, a SHIFT control is provided. This control is in the cathode circuit of the reactance-tube modulator and adjusts the amount of current through the valve and hence the amount of frequency shift. Since the reactance tube is only effective in the mark condition, this control has no effect on the space frequency. Consequently, it is adjusted for the correct mark frequency.

(iv) When F.S.-FAX switch is in the FAX position S_5 provides for photo-facsimile operation It is normally left in the F.S. position.

(v) The KEYING POLARITY switch S_6 provides for the various types of keying signal. It is adjusted so that the transmitter radiates on the high side of the assigned carrier while the keyboard is producing mark impulses.

(vi) The OUTPUT LEVEL control, a screwdriver-adjusted potentiometer R_{15}, is located at the rear of the unit. It adjusts the output to the value required to drive the transmitter by varying the screen voltage on V_6, the power amplifier.

(vii) The LINE-MARK-SPACE switch S_4 is used during the initial setting up of the keyer. By means of this switch the equipment may be placed in either the mark or space condition independent of keying. In the LINE position, the keyer valve is controlled by the keyboard.

(viii) C_{21}, a screwdriver adjustment accessible from the front panel is preset, but may be adjusted if it is found that the frequency control does not provide the correct adjustment of the 200 Kc/s oscillator frequency.

(ix) C_{33}, a preset capacitor in the phase-shifting network is factory set and normally does not require adjustment.

(x) The MAIN POWER switch S_1 completes the mains supply to T_1 in the power supply.

(xi) The PLATE ON-OFF switch S_2 removes the H.T. supply from the crystal oscillator, balanced modu-

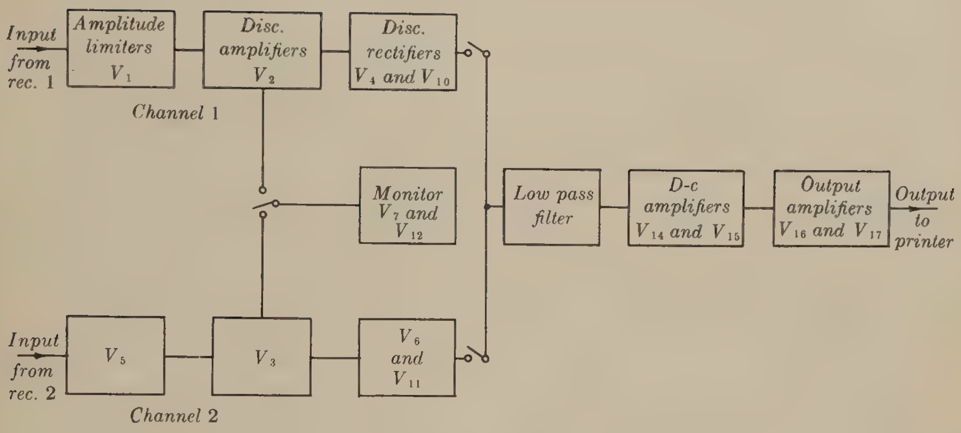


FIG. 19-18 Block diagram of frequency-shift converter.

lator and power amplifier for the stand-by position.

(xii) R_7 is a screwdriver-adjusted potentiometer in the screen circuits of the balanced modulator to compensate for inherent differences in the amplifications of the valves. It is adjusted to produce maximum cancellation of the crystal frequency in the balanced modulator tank circuit.

(xiii) Plug P_1 connects the output from the keyer to any one of three transmitters which may be permanently connected to the keyer.

19.8 Frequency-shift converters.

For long-distance sky-wave transmission, *space-diversity reception* is generally required to maintain reliable operation of radio teletype equipment. In this system, two or more separate receiving aerials are connected to individual radio receivers, the outputs of which are suitably combined. The aerials should be separated several wavelengths apart to be effective. Under these circumstances, the sky-wave signals arriving at the two aerials usually fade independently of each other and an input signal to the converter unit is maintained. An advantage of 10 to 15 *db* is usually obtained with space diversity reception as compared with single aerial operation. If it is inconvenient to use spaced aerials required in space-diversity reception, *frequency-diversity* operation may be substituted. This involves sending the signal at two or more frequencies which are separately received and combined in the outputs of the receivers. Alternatively, *polarization-diversity* reception may be employed. This involves the use of two or more aerials so oriented as to receive differently polarized components of the r-f signal. Its operation is based on the fact that the vertical and horizontal components of a received sky wave do not usually fade simultaneously.

A. Block Diagram of Converter

A block diagram of a frequency-shift converter is shown in Fig. 19-18. The valve numbers refer to the circuit diagram shown in Figs. 19-26, 19-27, and 19-28. The converter is used to convert the audio output of the receiving circuits into pulses of current of sufficient amplitude to drive a teletypewriter. Two similar input channels are provided for diversity reception. If diversity operation is not used, either channel may be disconnected. The input to each channel is an audio-frequency signal which shifts in frequency from mark to space condi-

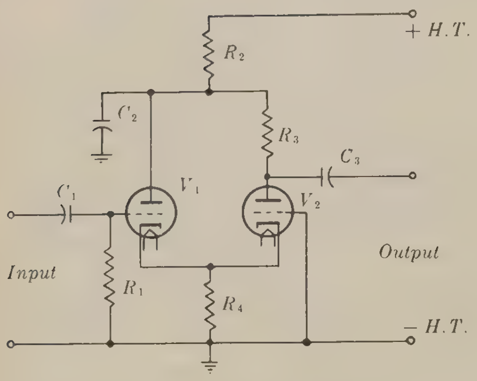


FIG. 19-19 Transient-free limiter.

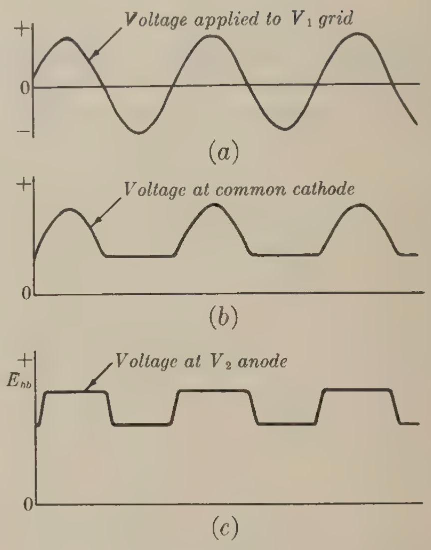


FIG. 19-20 Voltage waveforms of transient-free limiter.

tion by a fixed amount applied at the transmitter.

The input to each channel is applied through a band-pass filter which determines the final over-all bandwidth before demodulation. This filter is not shown on the block diagram. In Service equipment, the bandwidth of the filter is usually 1000 c/s. The signals are limited in amplitude and passed through a frequency-discriminating network to yield a d-c teletype signal. This signal is passed through a low-pass filter to remove carrier ripple and higher frequency noise components and is then amplified to a suitable level to operate automatic printing or recording apparatus.

1. LIMITER.

Since at this stage the carrier and intelligence frequencies are comparable, the transients in the limiter circuit must be extremely short to both frequencies. To meet these requirements, a *transient-free limiter* of the type shown in Fig. 19-19 is used. Consider the effect of a high-amplitude signal applied through capacitor C_1 to the grid of V_1. Small negative voltages cut off the valve and so the voltage across R_4 due to current through V_1 is zero during most of the negative half cycle. As the grid swings positive with respect to ground, current through V_1 increases, increasing the voltage across R_4 as shown in Fig. 19-20 (a). R_4 is made sufficiently large so that at no time does the grid voltage exceed the cathode voltage. Thus the grid never goes positive with respect to its cathode and no charge appears on C_1 which would have to leak off through R_1 in accordance with the time constant R_1C_1. Therefore the circuit is instantaneous in its action and no transients result. Also the load presented to the preceding stage remains constant.

As the voltage across R_4 increases because of the positive swing of V_1 grid, V_2 is cut off. V_2 is essentially a cathode driven stage excited by V_1 acting as a cathode follower. The gain of this stage is low because of its small anode resistor R_3. When the input to V_1 swings a certain amount positive V_2 cuts off; when the input to V_1 swings the same amount negative V_1 cuts off. Therefore the action of the limiter is symmetrical about the zero axis, and is both transient-free and instantaneous for any abrupt changes in amplitude or frequency. The waveforms of voltage at the common cathode and at the anode of V_2 are

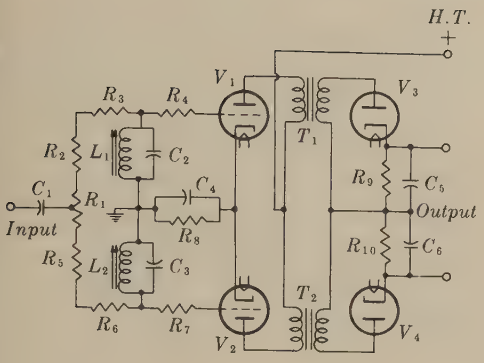


FIG. 19-21 Two-band-pass discriminator.

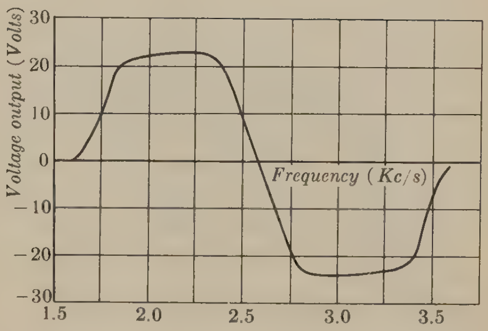


FIG. 19-22 Response curve of two-band-pass discriminator.

shown in Fig. 19-20 (b) and (c). To achieve the desired degree of limiting, two such limiters, joined by a Class *A* amplifier are used.

2. DISCRIMINATOR.

The discriminator is of the balanced type having two branches allowing differential combination of the two rectified outputs. This minimizes the response to amplitude modulation not eliminated by the filter. Two general types of discriminators are used. One consists of two band-pass filters centred about the mark and space frequencies respectively and effectively dividing the total band into halves. The other consists of a two-branch network, each branch of which has a varying amplitude characteristic extending over the complete transmission band and usually well beyond. The amplitude-frequency characteristics of these two branches have opposite slopes and are of such shapes that differential combination of their rectified outputs is an approximately linear voltage-frequency curve, passing through zero at mid-band.

A discriminator of the two-band-pass type is shown in Fig. 19-21. The input from the limiters is applied through the coupling capacitor C_1 to the grids of the discriminator amplifiers V_1 and V_2. Between grid and cathode of each amplifier is a parallel resonant circuit L_1C_2 and L_2C_3 for V_1 and V_2 respectively. One tank circuit is tuned to the mark frequency and the other to the lower space frequency.

V_1 and V_2 are fed in parallel via C_1. The input to each tube consists of the voltage developed across its respective tuned circuit. Considering V_1, the input to its grid is via C_1 in series with a portion of R_1, R_2, R_3 and L_1 and C_2 in parallel. At the frequency to which L_1 and C_2 are tuned, their impedance is maximum, and the voltage across them is maximum and appears at V_1 grid. At all other frequencies, L_1 and C_2 offer only a low impedance; a larger voltage appears across a part of R_1, R_2 and R_3, a low voltage is developed across L_1 and C_2 and at the grid of V_1. Thus the output of V_1 is a maximum at the frequency to which L_1 and C_2 are tuned, resulting in maximum voltage appearing across the primary of T_1, its anode load. Similarly, in V_2, the other part of R_1, R_5, R_6 and C_3 and L_2 discriminate against all frequencies other than that to which C_3 and L_2 are tuned. The discriminator rectifiers, V_3 and V_4, and their associated load resistors R_9 and R_{10} rectify the voltages applied to them. The load resistors are connected in opposition and by-passed by the capacitors C_5 and C_6. The two parallel resonant circuits L_1C_2 and L_2C_3 are tuned to the mark and

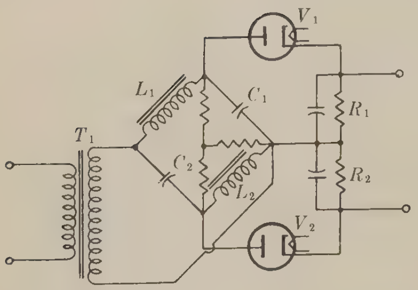


FIG. 19-23 Two-branch-network discriminator.

space frequencies; thus during the mark condition, the output voltage is of one polarity, the space of the opposite polarity and the reversal of the polarity occurs mid-way between the frequency extremities. The response of a discriminator of this type is shown in Fig. 19-22.

Shown in Fig. 19-23 is a discriminator of the two-branch-network type. Discriminator action is obtained by making use of the fact that if, in a series resonant circuit, a constant voltage is applied across the series inductor and capacitor, the voltage across the capacitor decreases with frequency and the voltage across the inductor increases with frequency. Across the secondary of the transformer T_1 is a bridge network consisting of two series resonant circuits in parallel, both of which are tuned to the cross-over frequency. V_1 is connected across C_1, the capacitor in the series resonant circuit composed of C_1 and L_1 while V_2 is connected across L_2, the inductor in the series resonant circuit L_2 and C_2. While the mark or higher frequency is being received, the frequency of the applied voltage will be higher than that to which the series resonant circuits are tuned. The voltage across the capacitor C_1 will be lower than that across the inductor L_2. V_2 will conduct more heavily than will V_1 resulting in a greater voltage being developed across resistor R_2 than R_1, the differentially-connected load resistors.

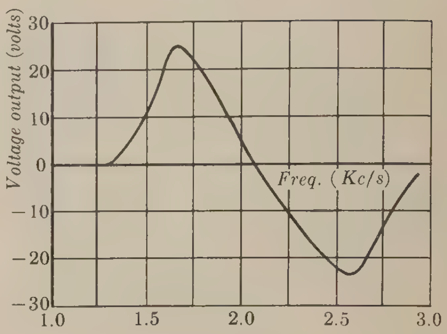


FIG. 19-24 Response curve of two-branch-network discriminator.

When the space frequency is being received, the voltage across C_1 will be greater than that across L_2, resulting in a greater voltage appearing across R_1 than R_2. The response curve of a discriminator of this type is shown in Fig. 19-24. Because of the almost linear relationship between frequency and voltage, this discriminator is sometimes referred to as the *linear discriminator*.

3. Low-Pass Filter.

The high frequency components in the demodulated output are filtered out by the low-pass filter if the noise level is low. As the noise level increases, there are short periods when the noise level exceeds the carrier. The action of the limiter is to give preference to the greater signal, in this case noise, and since the noise will appear to the discriminator as a carrier fluctuation around the mid-band as centre, the demodulator output momentarily dips toward zero. As the noise increases, the duration and frequency of these holes increase. The low-pass filter, by excluding frequencies considerably in excess of the maximum signalling speed, prevents these holes in the signal from producing false or extra transitions in the signal output. The low-pass filtering, however, cannot prevent the true transitions from being displaced by this type of noise component since the signal is obliterated momen-

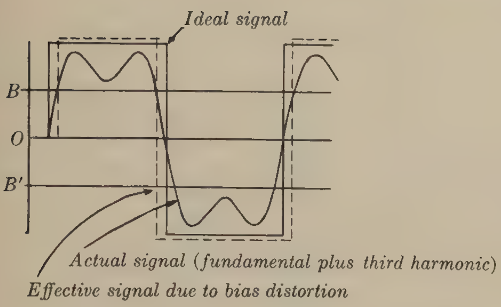


FIG. 19-25 Bias distortion resulting from change in d-c level.

tarily. Its most important function is to prevent a breakup in the signal until a fairly high distortion is reached.

4. D-C AMPLIFIER.

The output of the discriminator and low-pass filter is the re-created pulse that originally keyed the transmitter. This signal contains the fundamental and third harmonic of the keying signal. Thus the output of the filter will approach the shape illustrated in Fig. 19-25. The first stage of the d-c amplifier is a non-linear positive-feedback amplifier which has but two output conditions representing the mark and space. Transition between the two conditions is caused by reversals about the zero voltage axis. When the mark and space frequencies are equally separated from the discriminator cross-over frequency, the output of the first d-c amplifier stage will be a symmetrical square wave as shown by the solid line in Fig. 19-25, with the mark and space of equal length. A drift in frequency in any frequency-determining circuit causes a change in the position of this zero voltage axis, and were this change in the d-c level applied to the d-c amplifier which keys about the zero axis, it would stay in one condition longer than the other, resulting in an increase in the length of one element at the expense of the other, as shown by keying about the axis B in Fig. 19-25. This type of distortion is known as *bias distortion.* This effect is reduced to a tolerable minimum by blocking the d-c component from the d-c amplifier input, allowing only the useful signalling frequencies to effect its operation. Following the first stage, one or more push-pull stages are employed to amplify the signal to an amplitude sufficient to operate the printer.

B. Frequency-Shift Converter Circuits

As previously stated, it is customary practice to have the mark and space correspond to the high- and low-frequency excursions respectively. Similarly, it is customary to operate the converter and the associated receivers in such a manner that the high-frequency and low-frequency audio outputs of the receiver correspond to the mark and space. The mark LCR circuit of the discriminator is fixed tuned to 2975 c/s and the space LCR circuit is made variable to accept various degrees of frequency shift. For the reception of a transmission in which the mark is 850 c/s higher than the space transmission, the b.f.o. in the receiver is adjusted to oscillate at a frequency 2975 c/s higher than the receiver i-f for mark and 2125 c/s for space. This output is applied to the input terminals of two identical converter channels. Either of these channels, or both, may be used as required. The circuit diagram of one channel is shown in Fig. 19-26. It consists of a two-stage limiter and discriminator.

1. LIMITER.

From the input terminal the audio signal is applied to the grid of V_{1A}, the first of two limiter amplifiers contained in the double triode V_1. The limiters operate on the overdriven-amplifier principle, producing saturation on the positive halves of the input cycle and cut-off during the negative half cycles. Because a second limiter follows the first, any amplitude noise or amplitude

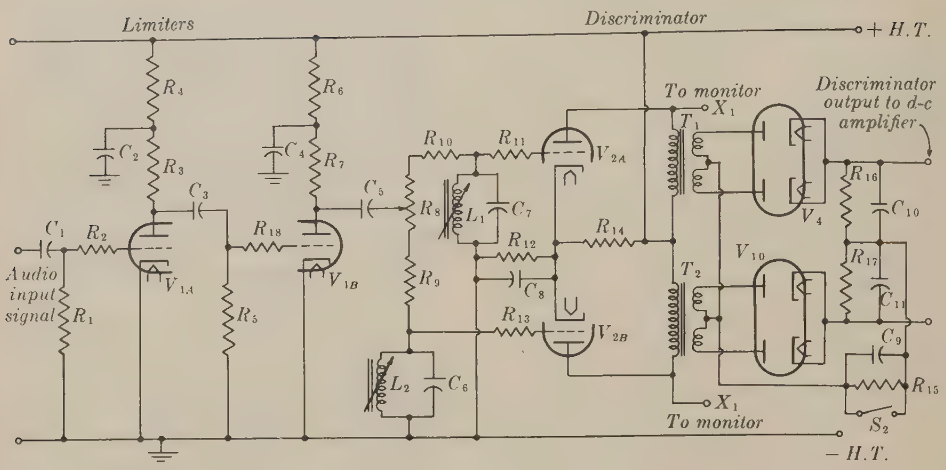


FIG. 19-26 Limiter and discriminator circuits.

modulation not removed by the first is removed by the second. The output of the limiters is essentially a square wave when the input is a sinusoidal voltage of sufficient amplitude.

2. DISCRIMINATOR.

The discriminator is of the two-band-pass type, in which V_{2A} and V_4 constitute the mark channel and V_{2B} and V_{10} form the space channel. Coil L_1 and capacitor C_7 are fixed tuned to 2975 c/s and L_2 and C_6 tuned to 2125 c/s for an 850 c/s shift. Resistors R_{16} and R_{17} form the differential-detector load, by-passed by C_{10} and C_{11}. R_{15} and C_9, a biasing network provided for on-off teletype are short-circuited by S_2 in the F.S. position.

3. PHASE INVERTER.

The low-pass filter, phase inverter, d-c amplifier and output amplifier circuits are shown in Fig. 19-27. The discriminator output, which consists of a series of positive and negative pulses (about 200 volts peak-to-peak for an 850 c/s shift) recording the intelligence, are impressed (via the low-pass filter consisting of C_1, C_2, L_1 and R_{11}) between grid and cathode of the left half of V_{14} (Fig. 19-27). The two halves of V_{14} act as an amplifier and phase inverter. Circuit components are such that in the absence of a signal both halves of the valve conduct. With the application of a signal, say positive to the grid of the left half of V_{14}, the anode current in this section of the valve increases resulting in a decrease in the anode voltage. The anode of the left half is coupled through R_6 to the grid of the right half. Therefore an increase in the current in the left half produces a decrease in current through the right half. This results in an increase in the anode voltage of the right half of the valve. Positive feedback is achieved by direct coupling from the anode of the right half to the grid of the left half via R_5. The action is cumulative to the point where high current flows in the left half and very low current flows in the right half. When the input changes polarity, to the space condition, the reverse of the above action occurs, with the left half conducting lightly, the right heavily. The change from one condition to the other is extremely rapid, producing essentially a square-wave push-pull output at the anodes of the left and right halves of the valve.

Because the voltage change required to initiate the above action is small,

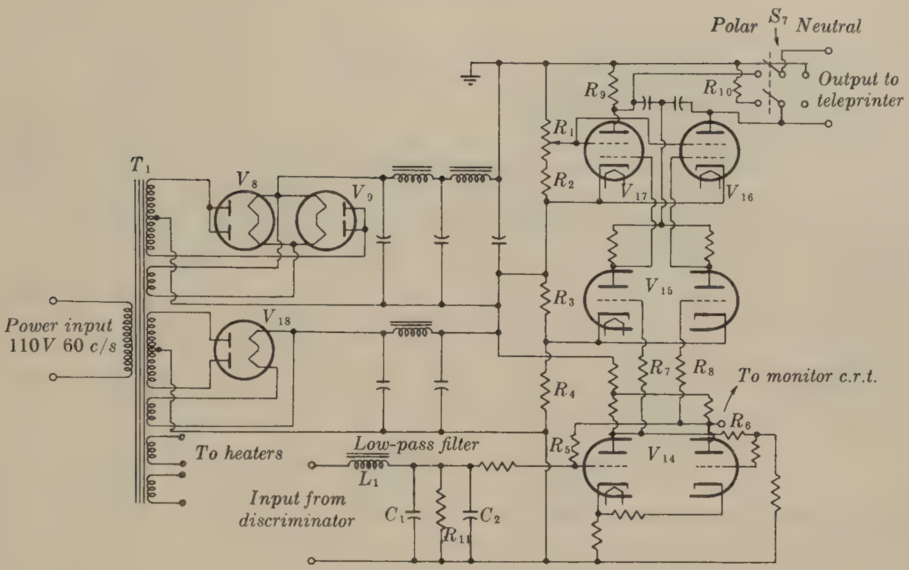


FIG.19-27 Low-pass filter, phase inverter, d-c amplifier and power supply circuits.

about .25 volts, compared to that supplied by the discriminator, a large degree of limiting exists. For this reason almost 100 volts of noise peaks may be tolerated before the noise can affect the output during the mark or space interval. During the transition between marks and spaces, the only effect noise can produce is to change the duration of a pulse by a small percentage.

4. D-C AMPLIFIER.

The double triode V_{15} comprises a conventional d-c amplifier. The cathodes are biased positively by the bleeder consisting of R_3 and R_4, the valves being alternately driven to cut-off and saturation by the output of V_{14}. The 1-megohm current-limiting resistors R_7 and R_8 prevent excessive grid current flow when the grids are raised above cathode potential.

5. OUTPUT STAGE.

Direct coupling is provided between the d-c amplifier and the output valves V_{16} and V_{17}. To provide for the relatively heavy current required to operate a teletypewriter two beam power amplifiers connected in push-pull are used in the output circuit. The input from V_{15} is of sufficient amplitude to drive the valves successively to anode current cut-off and to a conducting condition in excess of $60mA$ controlled by the adjustment of R_1 in the screen circuit. No grid-current limiting resistors are required in this stage because the method of coupling makes it impossible for the grids to be driven positive to the cathodes. Thus one valve is conducting while the other is cut off. Current is supplied to the line through the conducting valve, the polar-neutral switch S_7 and ground circuit. Consider first the action of the circuit with S_7 in the left hand position, that for polar operation. R_9 and the teletypewriter circuit are placed in parallel with the anode load resistor R_{10} so that when V_{16} conducts, some current flows through each of these branches. When V_{17} conducts, R_{10} and the teletypewriter circuit are effectively in parallel with R_9 and current flows through the teletypewriter line in the opposite direction to that current from V_{16}. In this way V_{16} and V_{17} supply current pulses to the

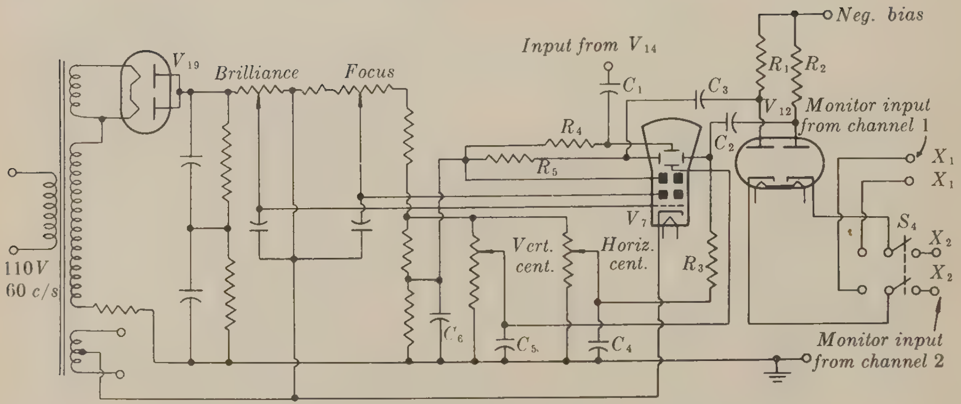


FIG. 19-28 Monitor circuit and power supply.

teletypewriter for mark and space polar operation.

With S_7 moved to the right, the position for neutral operation, V_{16} provides all the current for the printer; V_{17} although not contributing directly to the printer operation, draws current when V_{16} is cut off to maintain a nearly constant drain on the power supply. Resistor R_{10} is entirely removed from the circuit and the load for V_{16} in this case is the line and the associated printer.

6. MONITOR.

The cathode-ray tube V_7 shown with its associated circuits in Fig. 19-28, is used to indicate visually the quality of the signal and to show when the receiving system is properly tuned to the signal. The voltage waveform appearing at the anode of the right half of V_{14} (Fig. 19-27) consists of a series of positive and negative pulses corresponding to the marks and spaces as shown idealized in Fig. 19-29 (a). These pulses are differentiated by C_1 and R_4. C_6 effectively grounds the lower end of R_4 at the frequency of operation. The sharp positive and negative pulses shown in Fig. 19-29 (b) appearing across R_4 are applied to the upper deflection plate of the c.r.t. If the system is adjusted to produce a mark on the 2925 c/s input, the change of voltage across R_4 will be in the positive direction moving the spot of light rapidly upward on the screen and permitting it to fall at the rate of the exponential decrease of voltage. Horizontal-deflection voltage is obtained from the discriminator amplifier selected by S_4. When this switch is placed in the left hand position, channel 1 is monitored. At the instant of transition from the space to mark condition, the spot on the c.r.t. moves rapidly from the centre of the screen to the upper limit of its travel and returns at an exponential rate to the centre. At the same time, an a-c voltage of 2925 c/s appears at the anode of V_{2A}, alternating symmetrically about the zero or earth potential, and is applied to the cathode of the left hand side of V_{12}, a double-diode rectifier. The anodes of V_{12} are biased negatively by the power supply so that the left hand side of V_{12} does not conduct until its cathode is driven sufficiently negative to overcome this bias. When the left side of V_{12} does conduct, negative pulses, somewhat less than half a cycle in duration, as shown in Fig. 19-29 (c), appear across R_1 and are applied via C_3 across R_5 to the left X-plate of the c.r.t. The resulting pattern of this combination of vertical- and horizontal-deflection voltages is shown in Fig. 19-29 (d). In spite

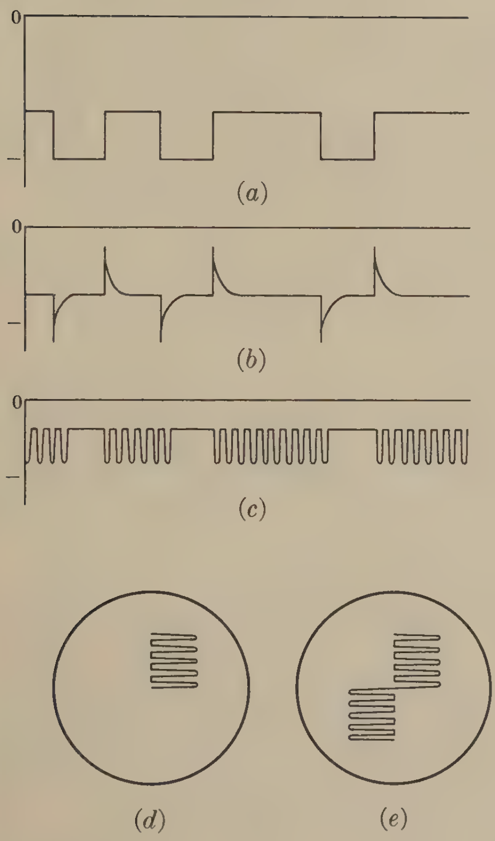


FIG. 19-29 Voltage waveforms in monitor circuit (a) Anode voltage of left half of V_{14} (b) Differentiated pulses from anode of V_{14} (c) Negative pulses applied to left X-plate (d) C.r.t. pattern resulting from space to mark transition (e) Complete monitor pattern for series of mark and space signals.

of the low response of V_{2B} (Fig. 19-26) to the mark frequency there is some output from this valve and were it not for the biasing of V_{12} this output could appear on the c.r.t. The biasing of V_{12} is sufficient that this does not occur during normal operation. In the transition from the mark condition to the space condition, the exact opposite of the above sequence occurs. The anode voltage of the right half of V_{14} drops sharply. This in conjunction with C_1 and R_4 produces a downward deflection on the c.r.t. Horizontal-deflection voltage is now obtained from V_{2B} via the limiter, which is the right hand section of V_{12}. The resulting pattern for a series of marks and spaces is shown in Fig. 19-29 (e). The fundamental frequency of the pulses produced in the five-impulse code is approximately 23 c/s. Thus the pattern on the tube during the reception of a steady signal consists of two *flags,* the 23 c/s ripple being barely perceptible. When the system is at rest, the pattern consists of a straight line extending from the centre of the tube to the right or left depending on whether it is resting in the mark- or space-frequency condition.

7. POWER SUPPLIES.

Two requirements essential in teletype operation necessitate an elaborate power supply arrangement. Since the lines connecting the output to the printer and the printer itself are not designed to carry high voltages, the output stage must operate at or near earth potential. Also, because the system must be able to stay in either the mark or space condition indefinitely, direct coupling must be used between stages in the pulse circuits. The first requirement is accomplished by grounding the positive side of the power supply; the second by using two full-wave rectifier power supplies in series. The d-c amplifier stage with the power supplies is shown in Fig. 19-27. R_1, R_2, R_3 and R_4 form a voltage divider tapped at points which maintain the proper voltages between the directly coupled stages. The monitor c.r.t. power requirements are met by a separate transformer and half-wave rectifier. Because of the low current drain, a simple RC filter is adequate. All power transformers are provided with adjustable taps enabling them to operate from either 110 or 220 volts, 50 or 60 c/s supply lines.

19.9 Misprinting. Misprinting in a radio teletype receiving installation can be attributed to three basic types of distortion. Fig. 19-30 (a) shows a

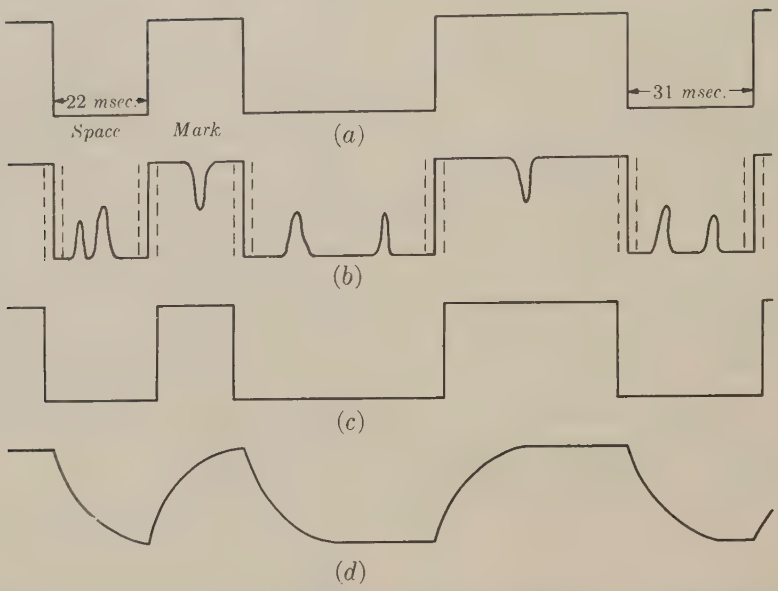


FIG. 19-30 Various types of distortion which may cause misprinting (a) Ideal signal waveform (b) Radio distortion (c) Bias distortion (d) Characteristic distortion.

perfect signal of the five-unit type, each unit being of 22 milliseconds duration. Radio distortion, illustrated by Fig. 19-30 (b), is characterized by random time shifts of the edges of the signal elements. Breaks and indentations of the signal shapes may also occur due to irregularities of radio transmission and other fortuitous fluctuations of the communications system. Fig. 19-30 (c) shows *bias distortion*, which is the lengthening of the space signal at the expense of the mark signal, or vice versa. Fig. 19-30 (d) shows *characteristic distortion*, that is, departure from the vertical in the leading edges and trailing edges of the pulses.

A. Radio Distortion

This subject along with its countermeasures is dealt with in preceding chapters. However, since the receiver output in radio teletype is in the form of two audio notes, the filtering in the audio system requires the special attention previously described. It must be wide enough to tolerate possible transmitter or receiver frequency drift. Noise signals outside the band are not completely eliminated, but due to the filter transients are re-introduced as signals of much lower amplitude.

In addition to band-pass filtering in the audio channel of the system, low-pass filtering of the discriminator output can attenuate noises of frequencies higher that that required to produce false printing. The circuits following the low-pass filter are designed to operate with an input consisting of a fundamental and third harmonic. An upper limit of 100 c/s is a common figure in Service equipment.

B. Bias Distortion

Bias distortion, by definition, is the lengthening of the mark or space signal at the expense of the other. It is expressed in percentage as

$$\frac{\text{mark length} - \text{space length}}{\text{mark length} + \text{space length}} \times 100$$

For an acceptable minimum number of errors, teletype signals can contain no more than 5% bias. Distortion of this

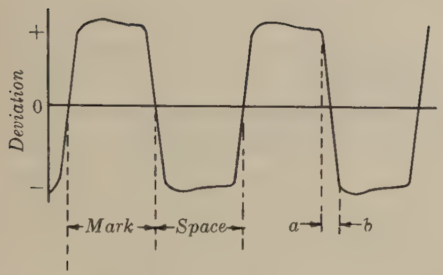


FIG. 19-31 Rounded waveforms which produce bias distortion.

type, provided that the transmitted signal is undistorted, originates in the converter stages. Fig. 19-31 shows how the sides of the waveform are sloped and the corners rounded off in the interest of reducing high-frequency transient radiations. The transition from the mark to the space frequency is not instantaneous, but requires a finite period of time designated by the distance *ab* on the diagram. Ideally, the transition from one condition to the other should occur midway between *a* and *b*. Failure in this respect produces bias distortion and results from two causes.

The first condition is illustrated in Fig. 19-32. If either the local oscillator or the b.f.o. frequency changes, the audio frequencies corresponding to the marks and spaces will also change, resulting in an output from the discriminator which is not symmetrically centred about the zero level. Although the d-c component of this voltage is blocked, the reversal of polarity occurs at a point on the wave where the length of one element is shorter than the other. Thus, for the conditions illustrated in Fig. 19-32 (a), the converter would produce a signal in which the spaces are longer than the marks as shown in Fig. 19-32 (b). A limited control over this type of distortion can be effected by discriminator tuning and bias adjustment in the d-c amplifier; however, its prevention rests primarily in a high degree of stability in all the frequency-determining circuits in the system.

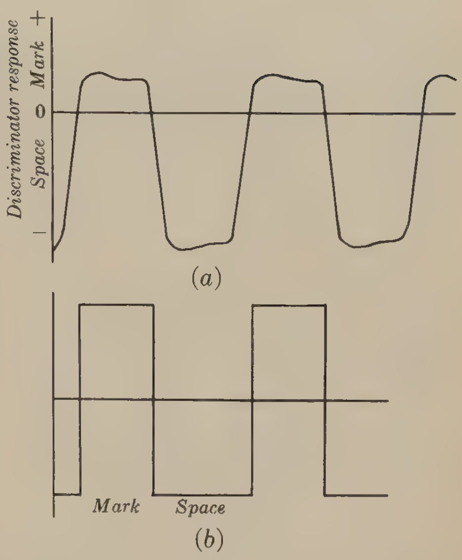


FIG. 19-32 Bias distortion resulting from frequency drift.

The second source of bias distortion is in the circuits following the discriminator. The first stage of these circuits must be extremely sensitive to changes in the polarity of the discriminator output, positive feedback being utilized to achieve this end. Also, it must be equally responsive to changes of polarity in both directions. The ideal condition is achieved when the stage following the discriminator changes from the mark to the space as the discriminator output passes through zero as shown by the solid lines in Fig. 19-33 (a) and (b). When the response of the circuit is not symmetrical, that is when it changes from one condition to the other at a voltage other than zero, as shown at *a* and *b* in figure 19-33 (a), distortion results as shown by the dotted lines in Fig. 19-33 (b).

Component tolerances are such that it is difficult to assemble a circuit which will meet the stringent requirements outlined above, and impossible to design

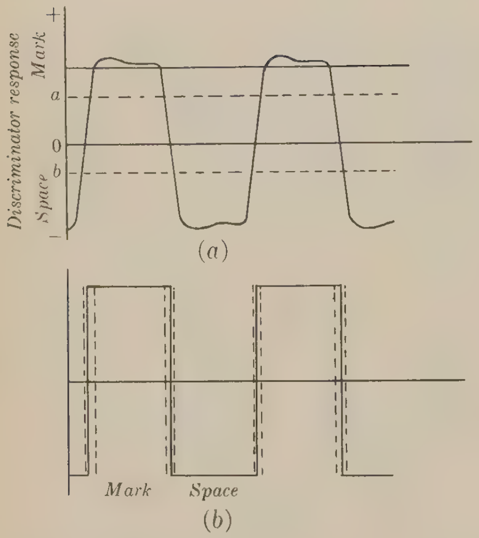


FIG. 19-33 Bias distortion resulting from asymmetric operation of circuit following limiter.

one which will continue to meet them. To compensate for these inaccuracies, the circuits include a variable *balancing* component to maintain bias distortion at an acceptable minimum.

The circuit following the discriminator provides sufficient amplitude to overdrive the succeeding stage. The sloping sides of the impulses are reduced to a negligible degree, thereby eliminating the possibility of further distortion of this type.

C. Characteristic Distortion

Characteristic distortion is the result of reactances in relay solenoids and transmission lines and is dealt with in Chapter XVIII.

19.10 Regeneration. It has been found that, disregarding the human element, but including propagation conditions and technical efficiency of the equipment, radio teletype should be capable of giving an efficiency factor of .85. However, due to inherent mechanical, electrical and propagation conditions, it becomes increasingly difficult to reach a unity factor by this means of communication. This factor can be approached to an estimated .98 by reducing all mechanical and electrical distortion to zero, leaving only the propagation conditions to introduce distortion into the system. The propagation conditions cannot be remedied, but the electrical and mechanical distortion in automatic radio reception can be effectively eliminated by the use of a *regenerator.*

The regenerator is fitted between the radio teletype converter and the teletype machine and automatically injects correctly timed pulses to take the place of those distorted by mechanical, electrical or propagation conditions, provided that a small part of the original pulse is evident. It is estimated that a completely faded signal under multi-path conditions should occur approximately 2% of the entire communication period, providing the correct frequencies are chosen, bringing the reception factor to approximately .98.

Two types of regenerators are in use, first the regenerative repeater whose operation is largely mechanical, and second the electronic regenerator. In radio teletype installations in the Service the electronic type is used.

A. Block Diagram

A block diagram of an electronic regenerator is shown in Fig. 19-34. This regenerator is specifically designed to correct for bias distortion in teletype signals and, when correctly adjusted, mark and space bias distortions of 45% are corrected to less than 5% at the output. While the regenerator compensates for considerable distortion in teletype links, there are some conditions which may cause faulty operation of the regenerator itself. The chief causes of faulty operation of the regenerator are listed below.

(i) If there is a missing start pulse due to fading of a signal, the regenerator

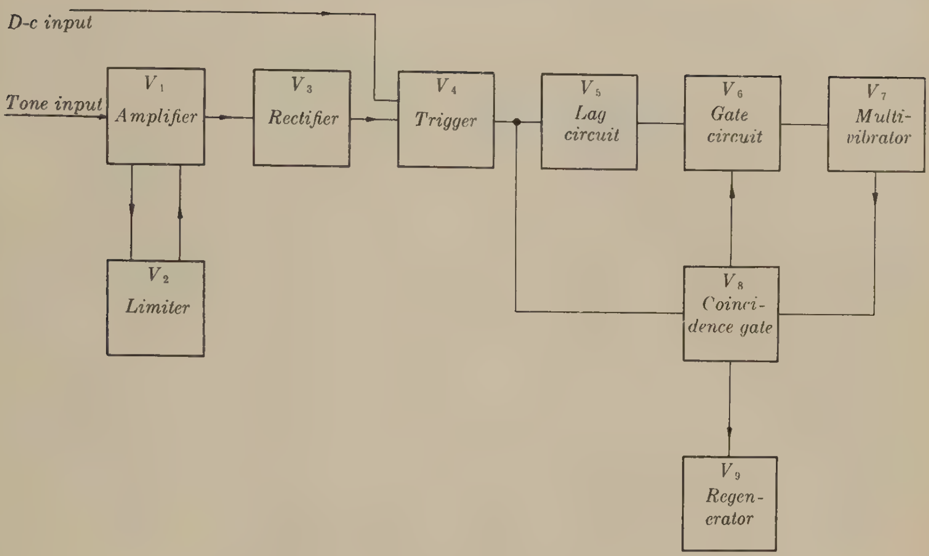


FIG.19-34 Block diagram of electronic regenerator.

will be started by the following intelligence pulse and will print errors for a few characters until the regenerator corrects itself.

(ii) If the start pulse is too short in duration (less than 10 msec. for 60 words per minute) it will be treated as noise and intercepted by the regenerator so that a following intelligence pulse longer than 10 msec. will act as the start pulse and errors will be printed for a few characters.

(iii) If there is a missing intelligence pulse, or if there is a split or knockout in one of the intelligence pulses at the point where sampling occurs, an error will be printed.

(iv) If the speed of the incoming signal changes, the amount of bias distortion that can be accepted by the regenerator will be reduced. Since the change in speed would appear as an increase in distortion to the operator, provision is made on the regenerator for the operator to perform periodic speed checks.

The electronic regenerator is intended to perform two general functions. The first is to regenerate single-channel teletype signals before operating a teletypewriter. The second is to regenerate diplex-teletype signals to single channel form for the operation of a teletypewriter.

Diplex operation is a method of doubling the traffic capacity of a teletype circuit. The two messages to be transmitted are tape-recorded and applied simultaneously to the single channel. The normal start and stop pulses are transmitted but each of the five selecting pulses is divided in halves. The first half of each pulse of the five normal pulses carries the intelligence of one tape recording, while the second half carries the intelligence of the other. At the receiving terminal the received intelligence is again recorded by two tape reperforators which are kept in synchronism by the start and stop pulses. One recorder is oriented to receive the first half-pulses while the other is oriented to receive the second half-pulses. The traffic capacity is doubled, for example, from a normal 60 to 120 words per minute.

In addition to normal pulse types of

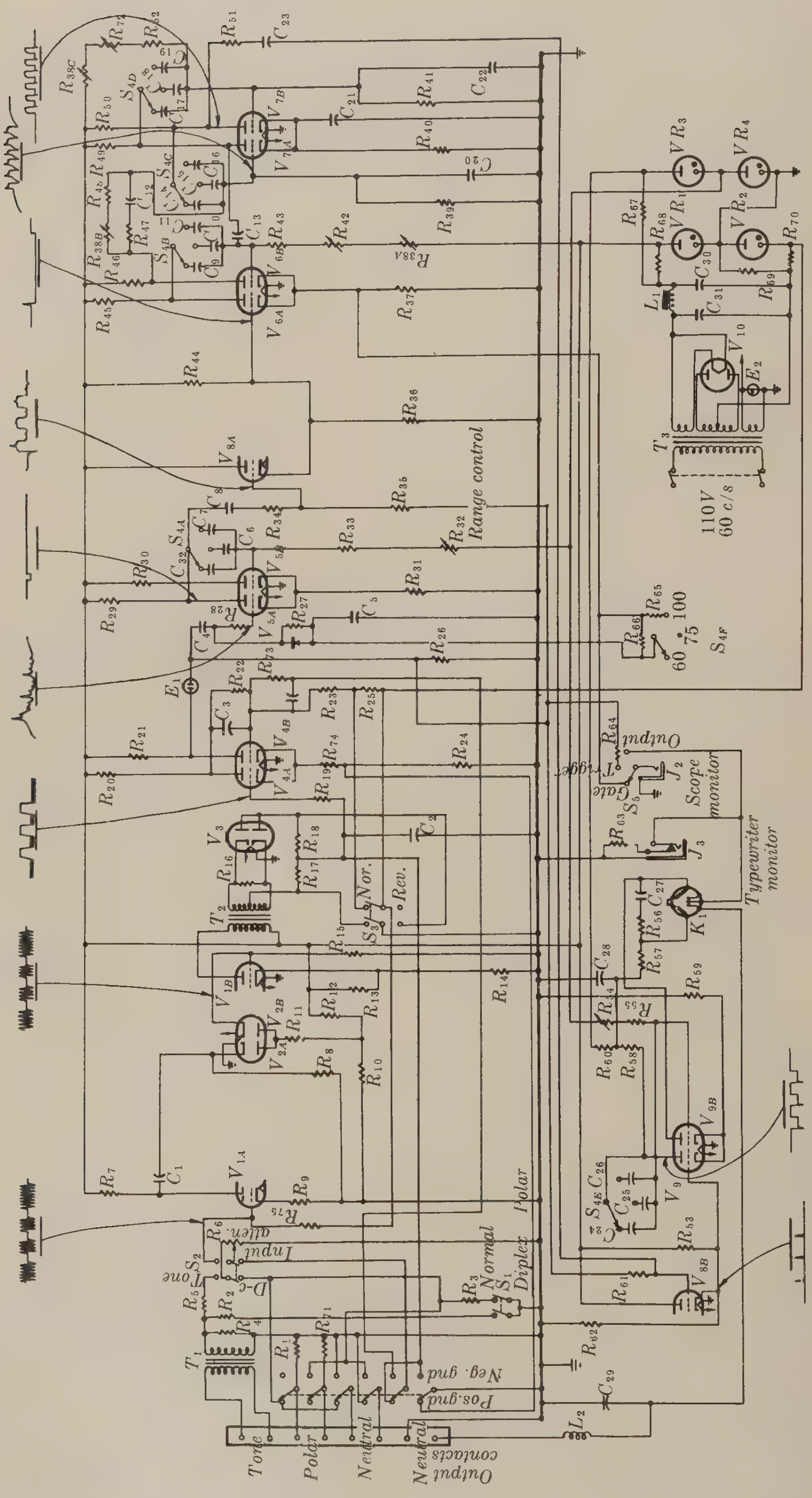


FIG. 19-35 Electronic regenerator circuit.

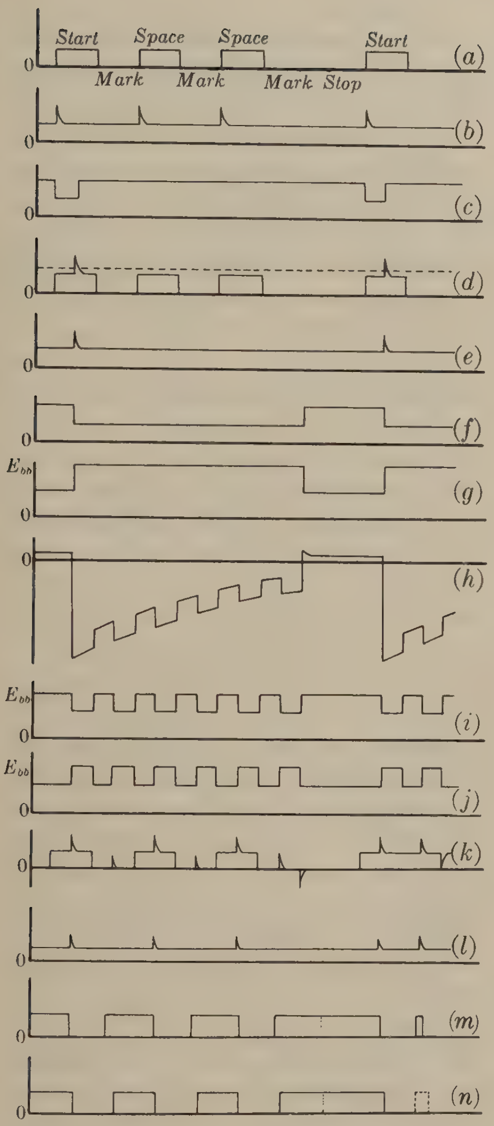


FIG. 19-36 Waveforms in electronic regenerator circuit (a) Teletypewriter signals (b) Voltage to grid of V_{5A} (c) Voltage at anode of V_{5A} (d) Voltage applied to grid of V_{8A} (e) Voltage at cathode of V_{8A} (f) Voltage at cathode of V_6 (g) Voltage at anode of V_{6B} (h) Voltage at control grid of V_{6B} (i) Voltage at anode of V_{7A} (j) Voltage at anode of V_{7B} (k) Voltage at grid of V_{8B} (l) Voltage at cathode of V_{8B} (m) Anode current of V_{9B} to solenoid (n) Line current through relay contacts.

mark and space teletype signals, provision is made for the application of keyed tone operation. In this type of operation the mark and space conditions are indicated by the presence or absence of an audio tone or vice versa. Although this method is not normally used in radio teletype operation, it is frequently used in land-line systems.

B. Regenerator Circuit

Fig. 19-35 is the circuit diagram of the regenerator; reference should be made to it and Fig. 19-36 waveforms throughout the following discussion.

1. INPUT AND AMPLIFIER CIRCUITS.

V_1, V_2 and V_3 are utilized in the amplifier to secure a d-c voltage to operate the following trigger circuit from tone inputs. The d-c input circuits present suitable impedances to the d-c input lines.

2. TONE INPUT.

Switch S_7 is placed in the positive ground position for all tone input keying. Keyed tone input to the regenerator is applied to the primary of the input transformer T_1 which has a nominal impedance of 600 ohms. Resistors R_2, R_4 and R_5 form a fixed attenuating network at the secondary of T_1 to prevent overloading amplifier V_{1A}. Switch S_2, set to TONE connects the output of T_1 to the grid of V_{1A} via input attenuator R_6. For diplex operation, switch S_1 is thrown to DIPLEX thus disconnecting resistor R_2 from across the secondary of T_1. This reflects an impedance of 1200 ohms to the primary so that when two regenerators are connected in parallel for tone input, the impedance is again 600 ohms and the operating level of the regenerators is relatively unchanged from that for simplex operation. V_{1A} and V_{1B} provide amplification of the tone signal so that low-level input signals will still operate the regenerator. Output from V_{1A} is capacitively coupled through capacitor C_1 to the series limiter V_2. V_2 is a dual diode series limiter which provides relatively uniform output with changing input levels. V_2 operates to clip the top off both the positive and negative halves of the sine wave input at a predetermined level as follows. The anodes of V_2 are returned to a slightly positive point (junction of R_{10} and R_{12}) through resistor R_{11} so that current flows in both sections of the valve. Signals of relatively small amplitude applied to the input cathode of V_2 are then passed to the output cathode without clipping. As the signal level is increased, a certain amplitude is reached where the voltage on the positive peaks of the sine wave applied to the input-cathode exceeds the static anode-to-cathode voltage. When this amplitude is reached, the valve ceases to conduct and clipping on the positive half of the wave results. The voltage developed across R_{11} is applied via V_{2B} to the grid of V_{1B}. On the negative half cycle the anode of V_{2B} becomes negative to its cathode causing V_{2B} to be cut off. This produces squaring of the negative half of the cycle at the grid of V_{1B}. By proper selection of R_8, R_{11} and R_{15} symmetrical clipping of the positive and negative halves of the signal is achieved. Output from the limiter V_2 is further amplified by triode V_{1B} which is transformer coupled to V_3. V_3 is a full-wave rectifier whose function is to supply a d-c voltage to operate the following trigger valve V_4. Resistor R_{16} loads the secondary of T_2 to reduce phase shift through the transformer. Keying switch S_3 is connected so as to secure either a positive or negative output from V_3. In the normal position of S_3 the centre tap of transformer T_2 secondary is grounded and a negative output at the junction of R_{17} and R_{18} is secured to operate the trigger valve. In the reverse position of S_3 the anodes of V_3 are grounded and a positive output at the junction of R_{17} and R_{18} is secured. Capacitor C_2 provides filtering of the rectified signal.

3. D-C INPUT.

When polar or neutral d-c keying is applied to the regenerator instead of tone input, switch S_2 must be set to D-C and switch S_7 located in the rear of the chassis is set to its proper position for positive or negative ground. The NORMAL-REVERSE switch S_3 should be in NORMAL for all d-c keying. The input attenuator resistor R_6 is then connected between the input line and the input to trigger valve V_4. For neutral simplex operation switch S_1 in the NORMAL position connects resistor R_3 in parallel with resistor R_1 across the input terminals to present an impedance of 310 ohms. For diplex neutral operation the d-c inputs of two repeaters are connected in parallel and switch S_1 is set to DIPLEX disconnecting resistor R_3. The input impedance of each unit is now 620 ohms and the paralleled impedance is 310 ohms. For polar simplex operation, resistors R_1 and R_{71} present a 1240 ohm input impedance balanced to ground. When two regenerators are connected in parallel, the polar input impedance is 620 ohms balanced to ground. Input attenuator R_6 provides a means of adjusting the level to the following trigger valve V_4 for either tone or d-c input to the regenerator.

4. TRIGGER CIRCUIT.

The trigger circuit is provided to accept the d-c input signals from either the rectifier valve V_3 or directly from the d-c inputs and provides a stable square wave teletype signal voltage at the trigger output. A uniform square voltage is essential for optimum operation of the following sampling pulse generating circuits and gating circuits. Trigger valve V_4 utilizes a direct-coupled flip-flop circuit to secure a square wave voltage across the output load resistor R_{26}. Threshold action is secured in V_4 by a negative bias at the grid of V_{4B} so that random input voltages below the level necessary to trigger the valve will have no effect on the circuit.

Keying switch S_3 shorts out resistor R_{25} when set for normal keying. This puts the grid of V_{4B} at a negative potential, no anode current flows and the neon lamp E_1 is on. Voltage of about 18V is now developed across resistor R_{26}. When the signal is applied to the regenerator as during a stop mark, a negative voltage is applied to the grid of V_{4A}. This reduces the anode current causing the anode potential of V_{4A} and the grid of V_{4B} to swing positive. Anode current flows in V_{4B}, the anode voltage drops below the operating potential required for neon lamp E_1 and the voltage across R_{26} drops to zero. Since application of a negative voltage to the grid of V_{4A} causes an increase in the anode current of V_{4B}, the current through the common cathode resistor R_{24} increases, causing the cathode bias voltage to increase. The effect on V_{4A} is in the same direction as the application of the initial negative voltage to the grid so that a regenerative action results and a very steep drop off in voltage across R_{26} occurs. Upon removal of the negative voltage to the grid of V_{4A}, the circuit drops back to its original state very rapidly due to the regenerative cathode coupling. Thus neon lamp E_1 is turned on instantly. Capacitor C_3 speeds up the action of the trigger circuits so that the rise and decay time of the voltage developed across R_{26} is very small (less than 25 msec.)

When keying switch S_3 is set to reverse, resistor R_{25} is inserted in the grid circuit of V_{4B} so that V_{4B} normally draws current. The voltage at the anode of V_{4B} is below the operating potential of E_1 and no voltage is developed across R_{26}. Switch S_3 also connects a positive output voltage from rectifier valve V_3 to the grid of V_{4A}. The inverse tone input signal will now have tone during the start pulse so that a positive d-c voltage will be applied to the grid of V_{4A}. The anode current of V_{4A} will increase causing the voltage at the anode of V_{4A} and at the grid of V_{4B} to swing in a negative direction. Valve V_{4B} is thus biased off causing the voltage at the anode to swing positive. Neon lamp E_1 turns on and a voltage is developed across R_{26} as before. The keying switch S_3 thus provides the same output keying across R_{26} regardless of whether the tone input is normal or reverse. The output keying across R_{26} is reverse (on during the start period) and is used to operate the following circuit. When d-c keying is applied to the regenerator, the input may be either positive or negative on stop mark.

5. SAMPLING PULSE GENERATING CIRCUITS.

Six sampling pulses, each of 1 msec. duration, are required for sampling the six teletype pulses used to transmit each character. These pulses are generated and adjusted for timing in V_5, V_6 and V_7. V_5 and V_6 are one-shot multivibrators of the cathode-coupled type. V_7 is a free-running multivibrator when it is gated on by V_6. The sampling pulses are secured by differentiating the output of V_7. The speed control varying the frequency of V_7 determines the spacing between sampling pulses (nominally 22 msec. for a speed of 60 words per minute). Gate valve V_6 determines the number of sampling pulses. V_5 and its associated range control vary the position of the sampling pulses in respect to the teletype signal pulses.

6. LAG CIRCUIT.

The trigger output signal developed across load resistor R_{26} is differentiated by capacitor C_4 and resistor R_{27} to provide a positive pulse to trigger the grid of the lag valve V_{5A}. V_{5B} normally conducts current due to the connection of its grid through R_{32} and R_{33} to a positive voltage. V_{5A} is then biased off due to the voltage developed across the common cathode resistor R_{31}. When a positive pulse of sufficient amplitude is applied to the grid of V_{5A}, the circuit triggers into a state where V_{5A} conducts and V_{5B} does not. After a time interval dependent mainly on the capacitance of C_{32} and the resistance of R_{33}, the circuit triggers back to its normal state with V_{5B} conducting current. The voltage at the anode of V_{5A} drops to about 100 volts when the circuit is triggered, and suddenly shifts back to the $+H.T.$ value at the end of the cycle. The change of voltage in a positive direction is differentiated by capacitor C_8 and resistor R_{34} plus R_{35} to supply a trigger pulse of proper amplitude to trigger gate valve V_6. This trigger pulse is gated by V_{8A} before being applied to the grid of V_{6A}. Range control R_{32} is a variable resistor in the timing circuit of V_5 and thus functions to change the operating time of V_5. Thus a delay between the leading edge of the start pulse and the first sampling pulse is made adjustable from about 2 msec. to 20 msec. For operation at 75 and 100 words per minute switch S_{4A} connects appropriate values of capacitors (C_6 or C_7) into the timing circuit of V_5 so that operation of the range control is approximately the same for teletype signals at speeds of 60, 75 or 100 words per minute.

7. GATE CIRCUIT.

Gate valve V_6 utilizes a one-shot multivibrator circuit similar to that of V_5 except that its operating time is much longer. V_6 accomplishes two major functions.

(i) Multivibrator V_7 is turned on and off.

(ii) Lag valve V_5 is rendered inoperative until the following stop mark.

The operating time of V_7 is determined primarily by capacitor C_9 and resistors R_{43}, R_{42} and R_{38A}. For speeds of 75 or 100 words per minute switch S_{4B} connects capacitors C_{10} or C_{11} into the timing circuit of V_7 to provide proper operation of the circuit at different teletype speeds. Variable resistor R_{38A} is part of a three-gang variable-resistance speed control and smoothly varies the timing of V_7 about $\pm 10\%$. Sections R_{38B} and R_{38C} of the speed control vary the frequency of the multivibrator V_7 in unison with the timing of gate valve V_6 so that for any setting of the speed control only $5\frac{1}{2}$ cycles of the multivibrator occur. The speed control nominally provides for $\pm 10\%$ variations in speed for speeds of 60, 75 or 100 words per minute selected by switch S_4. Variable resistor R_{42} provides a fine adjustment of the timing of V_6 so that variable resistor R_{38A} tracks properly (gates only $5\frac{1}{2}$ cycles) with R_{38B} and R_{38C}. Multivibrator V_7 is turned on for $5\frac{1}{2}$ cycles as follows. V_7 utilizes a symmetrical multivibrator circuit with the exception of resistor R_{46} between R_{38B} and $+H.T.$ Neglecting resistor R_{46}, resistors R_{38B} and R_{39} form a divider network between $+H.T.$ and ground with the grid of V_{7A} connected to the junction of R_{48} and R_{39}. An equal resistor network from +H.T. to ground consisting of R_{38C}, R_{52} and R_{41} has the grid of V_{7B} connected to the junction of R_{52} and R_{41}. Equal anode resistors R_{49} and R_{50} and equal coupling capacitors C_{14} and C_{17} provide a symmetrical circuit which will operate at 45.45 cycles per second (nominal for 60 words per minute) if gate valve V_6 is out of the circuit. When gate valve V_6 is in the circuit, V_{6B} conducts current due to the connection of its grid to $+150$ volts through resistors R_{43}, R_{42} and R_{38A}. The voltage at the anode of V_{6B} is then below 100 volts due to the voltage drop across the anode resistor R_{46}. The voltage at the grid of V_{7A} is now considerably below the voltage at the grid of V_{7B}, since the resistor network connected to the grid of V_{7A} connects the grid to a positive potential of 22.5 volts, while the resistor network at the grid of V_{7B} connects it to a potential of 50 volts. Under these conditions V_{7B} conducts considerable current causing sufficient voltage drop across the common cathode resistor R_{40} to bias off V_{7A}, and V_7 does not oscillate. At the moment V_6 is triggered by a pulse from lag valve V_5, the output section of V_6 (V_{6B}) ceases to conduct and multivibrator V_7 starts to oscillate. To make sure that oscillations start immediately in V_7, capacitor C_{12} and resistor R_{47} apply a pulse from the anode of V_{6B} to the grid of V_{7A} at the moment of starting. Capacitor C_{13} applies positive pulses at each half-cycle point from the anode of V_{7A} to the grid of V_{6B}. At the time of the sixth positive pulse, the voltage at the grid of V_{6B} is nearly to the point where the valve will trigger back to its stable state with V_{6B} conducting. The sixth positive pulse causes this triggering action to occur so that V_6 always triggers off at the $5\frac{1}{2}$ cycle point of the multivibrator operation and in turn shuts off the multivibrator until the gate valve V_6 is again triggered on by the lag valve V_5. Capacitors C_{20}, C_{21} and C_{22} in the multivibrator circuit help to stabilize the frequency of oscillation. Capacitors C_{15} and C_{18} provide the proper frequency of operation of valve V_7 for a speed of 75 words per minute. Capacitors C_{16} and C_{19} provide the proper frequency of operation of V_7 for 100 words per minute.

Lag valve V_5 is rendered inoperative by gate valve V_6 after the start pulse in the following manner. The grid of V_{5A} is returned to the cathode of V_6 through resistors R_{28}, R_{27} and R_{66}. When the valve V_6 is in its off state (V_{6B} conducting current), a positive voltage is developed across the cathode resistor of V_5 so that the grid of V_{5A} is just biased off. Under these conditions, a positive pulse of normal amplitude is sufficient to trigger the grid of V_{5A}. When V_6 is triggered from V_5, V_{6A} conducts while V_{6B} does not. Since the anode resistor of V_{6A} is much larger in value ($R_{45} =$ 33,000 ohms) than the anode resistor of V_{6B} ($R_{46} =$ 10,000 ohms), the voltage across the cathode resistor R_{37} will be much less when gate valve V_6 is on than when it is off. Thus the grid of V_{5A} will be returned to a much more negative point with respect to its cathode when the gate is on than when the gate is off. Under these conditions the pulses following the start pulse, although applied to the grid of V_{5A}, are not of sufficient amplitude to trigger the circuit. About 20 msec. after the gate valve turns off (V_{6B} conducts current), the lag valve V_5 is again capable of being triggered by the first pulse applied to the grid. The 20 msec. delay is caused by capacitor C_5 and resistor R_{66}. At 100 words per minute this delay is reduced by shunting R_{66} by R_{65} to reduce the RC time constant. The purpose of the delay is to permit the regenerator to recover quickly if a start pulse is missing and the gate cycle is started by a following pulse.

8. Coincidence Gate Circuit.

V_{8A} performs the function of preventing random pulses of noise at the regenerator input from triggering the gate circuit providing the noise pulses are somewhat shorter than half the normal starting pulse duration. V_{8B} provides the circuit in which the teletype signal pulses are sampled by the one msec. pulses from the multivibrator and thus determines whether a mark or space is to be produced by the output of the regenerator.

V_{8A} is one half section of a double triode which is connected as a cathode follower. The valve is normally biased off by cathode resistor R_{36} and resistor

R_{44} connected between cathode and +H.T. The grid of V_{8A} is connected through resistor R_{35} to the gating voltage developed at the output of V_4 and also receives pulses from lag valve V_5 through capacitor C_8 and resistor R_{34}. At the time of the teletype start pulse the gating voltage of +17 volts is applied to the grid of V_{8A} so that the valve is biased just beyond cut-off. A pulse now applied through capacitor C_8 and resistor R_{34} to the grid will cause anode current to flow and the pulse will appear at the cathode. If the pulse is applied to the grid of V_{8A} when there is no gating voltage, V_{8A} will be biased so far beyond cut-off that the amplitude of the pulse will not be sufficient to cause anode current to flow and the pulse will not appear at the cathode. Assuming a random pulse of 8 msec. duration is applied to the regenerator, this pulse will cause a gating voltage of eight msec. duration to be applied to the grid of V_{8A}. The pulse to the grid of V_{8A} through capacitor C_8 under normal setting of the range control at 60 words per minute operation will not occur until 11 msec. after the start of the 8 msec. gating pulse. The pulse through capacitor C_8 will thus occur three msec. after the gating voltage has ceased and will not appear at the cathode of V_{8A} to trigger the following gate valve V_6. As long as V_6 is not triggered, no output can be secured from the regenerator.

9. SAMPLING GATE.

V_{8B} is connected in a similar circuit to that of V_{8A}. Cathode-to-ground resistor R_{62} and cathode-to-+H.T. resistor R_{53} bias V_{8B} far beyond cut-off. The output from the multivibrator V_7 is differentiated and applied to the grid of V_{8B} through capacitor C_{23}. Gating voltage to the grid of V_{8B} is applied through resistor R_{61} so that when the +17 volts is on, as during the start pulse, the positive differentiated multivibrator pulse causes anode current to flow in V_{8B} and a positive pulse appears at the cathode to drive the regenerator valve V_9. Resistor R_{61} prevents the differentiated multivibrator pulse from feeding back into the source of the gating voltage across resistor R_{26} with sufficient amplitude to cause any interlocking in operation. When there is no gating voltage applied to V_{8B}, the pulse from the multivibrator is not of sufficient amplitude to cause anode current to flow and the pulse does not appear at the cathode. Resistor R_{51} limits the amplitude of the multivibrator pulse to secure the desired operation.

10. REGENERATOR CIRCUIT.

The regenerator circuit provides the output teletype signals in normal keying form and with less than 5% distortion. The one-shot multivibrator regenerator circuit employs V_{9B} and is normally excited so that relay K_1 has its contacts closed during any steady input condition of the regenerator. When a teletype signal is applied to the regenerator, a positive pulse is impressed on the grid of V_{9A} from the cathode of the sampling gate valve V_{8B} every time a space pulse occurs. With normal adjustments for 60 words per minute, V_9 will trigger into its unstable state upon application of the positive trigger pulse and remain in this state for about 19 msec. before triggering back to the stable state. This operating time interval is determined primarily by the values of capacitor C_{24}, resistor R_{55} and R_{54}. Resistor R_{54} is the distortion control and permits adjustment of this time by at least ±10%. At the instant V_{9A} is triggered by the pulse from V_{8B}, V_{9B} ceases to conduct and the relay K_1 opens within 2 msec. Assuming there is no space pulse following the start pulse, the current will not be applied to the relay again for 19 msec. and the relay coil characteristics are such that the contacts will not close for five msec. after the current has been applied. The total time the relay contacts are open can be calculated from the following formula. Contact open time equals circuit operating time minus time for contacts to open plus time for contacts to close. The relay contact time thus can be evaluated at $19-2+5$ or 22 msec. which is the correct time of a space signal. A second pulse 22 msec. after the first pulse will trigger V_9 from the sampling gate valve V_{8B}. There will be a three msec. period when the relay coil is energized under these conditions (22 msec. between pulses to circuit minus 19 msec. operating time of circuit). However, since the relay requires five msec. to close after being energized, the contacts will not close at all and a 44 msec. space will result from the relay contacts. In this fashion the relay will produce undistorted teletype signals at the contacts which will depend only on the regenerator valve V_9 receiving the 1 msec. driving pulses from the sampling gate circuit. C_{25} and C_{26} are selected by speed switch S_{4E} to provide proper circuit timings at 75 and 100 word per minute operation. Capacitor C_{27} and resistor R_{56} determine to some extent the operating time characteristics of relay K_1. Capacitor C_{28} and resistor R_{60} form a decoupling network in the +H.T. supply to V_9. More uniform operation of the circuit is thereby achieved. Jack J_3 provides connection for a monitor teletypewriter. C_{29} and r-f choke L_2 prevent radiation of excessive noise voltages generated by the contacts of relay K_1.

19.11 Questions and problems.

1. State three advantages of teletype over telegraph as a means of communication.

2. Describe with the aid of a diagram the current variations in the teletype line required to transmit the letter *B*.

3. Explain briefly how the start-stop mechanism operates in the segmented-ring distributor when a key is struck.

4. Explain why a 10% variation in speed of the receiving brush assembly (Fig. 19-4) is unlikely to cause printing errors.

5. Describe briefly the function of each of the following components of a rotating-cam teletypewriter
   (a) universal bar
   (b) clutch
   (c) locking levers
   (d) lock-loop cam.

6. Describe the function of the sword of Fig. 19-9 during the reception of a mark impulse.

7. How does the use of perforated tapes add to the efficiency of teletype communication?

8. Explain the purpose of each of the following
   (a) perforator
   (b) reperforator.

9. Differentiate between chad and chadless tapes.

10. What is the function of a transmitter-distributor?

11. Express a speed of 45 words per minute in c/s and in bauds.

12. Why is frequency-shift keying preferable to other methods of modulation in radio teletype service?

13. What is the meaning of the term phantom carrier?

14. State the mark and space frequencies which would be radiated from a radio teletype transmitter operating with a phantom carrier frequency of 7685 Kc/s.

15. Make a block diagram to show the equipment required for
    (a) the transmitting terminal
    (b) the receiving terminal of a radio teletype link.

16. State the purpose of the keyer unit and of the converter unit in a radio teletype system.

17. Draw a block diagram of a communications receiver and indicate at what frequency each stage will be operating during the reception of mark and space signals, given the following frequencies
    (a) phantom carrier, 8.5 Mc/s
    (b) frequency swing, 850 c/s
    (c) i-f, 750 Kc/s, with local oscillator operating above the carrier frequency
    (d) audio notes of 2975 c/s for mark and 2125 c/s for space signals.

18. Explain how the frequency-shift keyer of Fig. 19-16 produces stabilized mark and space signals with a phantom carrier of 345 Kc/s.

19. What advantage is gained by the use of space-diversity reception?

20. Distinguish between space-diversity reception and frequency-diversity reception.

21. Draw a diagram and explain the action of a two-band-pass discriminator.

22. Why is direct coupling required in the amplifier stages of a radio teletype converter?

23. Explain the origin of bias distortion and characteristic distortion.

24. What is the effect of bias distortion on the operation of the teletypewriter?

25. How may bias distortion be corrected?

26. What is the purpose of a teletype regenerator?

27. Draw a block diagram of an electronic regenerator and state briefly the function of each component.

APPENDIX

LOGARITHMS

	0	1	2	3	4	5	6	7	8	9	Mean Differences								
											1	2	3	4	5	6	7	8	9
10	00000	00432	00860	01284	01703						42	85	127	170	212	254	297	339	381
						02119	02531	02938	03342	03743	40	81	121	162	202	242	283	323	364
11	04139	04532	04922	05308	05690						37	77	116	154	193	232	270	309	348
						06070	06446	06819	07188	07555	37	74	111	148	185	222	259	296	333
12	07918	08279	08636	08991	09342						36	71	106	142	177	213	248	284	319
						09691	10037	10380	10721	11059	34	68	102	136	170	204	238	272	307
13	11394	11727	12057	12385	12710						33	66	98	131	164	197	229	262	295
						13033	13354	13672	13988	14301	32	63	95	126	158	190	221	253	284
14	14613	14922	15229	15534	15836						30	61	91	122	152	183	213	244	274
						16137	16435	16732	17026	17319	29	59	88	118	147	177	206	236	265
15	17609	17898	18184	18469	18752						28	57	85	114	142	171	199	228	256
						19033	19312	19590	19866	20140	28	55	83	110	138	165	193	221	248
16	20412	20683	20951	21219	21484						27	53	80	107	134	160	187	214	240
						21748	22011	22272	22531	22789	26	52	78	104	130	156	182	208	233
17	23045	23300	23553	23805	24055						26	50	76	101	126	151	176	201	227
						24304	24551	24797	25042	25285	25	49	73	98	122	147	171	196	220
18	25527	25768	26007	26245	26482						24	48	71	95	119	143	167	190	214
						26717	26951	27184	27416	27646	23	46	69	93	116	139	162	185	208
19	27875	28103	28330	28556	28780						23	45	68	90	113	135	158	180	203
						29003	29226	29447	29667	29885	22	44	66	88	110	132	154	176	198
20	30103	30320	30535	30750	30963	31175	31387	31597	31806	32015	21	43	64	85	106	127	148	170	190
21	32222	32428	32634	32838	33041	33244	33445	33646	33846	34044	20	40	60	80	100	120	140	160	180
22	34242	34439	34635	34830	35025	35218	35411	35603	35793	35984	20	39	58	77	97	116	135	154	174
23	36173	36361	36549	36736	36922	37107	37291	37475	37658	37840	19	37	56	74	93	111	130	148	167
24	38021	38202	38382	38561	38739	38917	39094	39270	39445	39620	18	35	53	71	89	106	124	142	159
25	39794	39967	40140	40312	40483	40654	40824	40993	41162	41330	17	34	51	68	85	102	119	136	153
26	41497	41664	41830	41996	42160	42325	42488	42651	42813	42975	16	33	49	66	82	98	115	131	148
27	43136	43297	43457	43616	43775	43933	44091	44248	44404	44560	16	32	47	63	79	95	111	126	142
28	44716	44871	45025	45179	45332	45484	45637	45788	45939	46090	15	30	46	61	76	91	107	122	137
29	46240	46389	46538	46687	46835	46982	47129	47276	47422	47567	15	29	44	59	74	88	103	118	132
30	47712	47857	48001	48144	48287	48430	48572	48714	48855	48996	14	29	43	57	72	86	100	114	129
31	49136	49276	49415	49554	49693	49831	49969	50106	50243	50379	14	28	41	55	69	83	97	110	124
32	50515	50650	50786	50920	51054	51188	51322	51455	51587	51720	13	27	40	54	67	80	94	107	121
33	51851	51983	52114	52244	52375	52504	52634	52763	52892	53020	13	26	39	52	65	78	91	104	117
34	53148	53275	53403	53529	53656	53782	53908	54033	54158	54283	13	25	38	50	63	76	88	101	113
35	54407	54531	54654	54777	54900	55023	55145	55267	55388	55509	12	24	37	49	61	73	85	98	110
36	55630	55751	55871	55991	56110	56229	56348	56467	56585	56703	12	24	36	48	60	71	83	95	107
37	56820	56937	57054	57171	57287	57403	57519	57634	57749	57864	12	23	35	46	58	70	81	93	104
38	57978	58092	58206	58320	58433	58546	58659	58771	58883	58995	11	23	34	45	57	68	79	90	102
39	59106	59218	59329	59439	59550	59660	59770	59879	59988	60097	11	22	33	44	55	66	77	88	99
40	60206	60314	60423	60531	60638	60746	60853	60959	61066	61172	11	22	33	44	55	66	77	88	99
41	61278	61384	61490	61595	61700	61805	61909	62014	62118	62221	10	21	31	42	53	63	74	84	95
42	62325	62428	62531	62634	62737	62839	62941	63043	63144	63246	10	20	31	41	51	61	71	82	92
43	63347	63448	63548	63649	63749	63849	63949	64048	64147	64246	10	20	30	40	50	60	70	80	90
44	64345	64444	64542	64640	64738	64836	64933	65031	65128	65225	10	20	29	39	49	59	68	78	88
45	65321	65418	65514	65610	65706	65801	65896	65992	66087	66181	10	19	29	38	48	57	67	76	86
46	66276	66370	66464	66558	66652	66745	66839	66932	67025	67117	9	19	28	37	47	56	65	74	84
47	67210	67302	67394	67486	67578	67669	67761	67852	67943	68034	9	18	27	36	46	55	64	73	82
48	68124	68215	68305	68395	68485	68574	68664	68753	68842	68931	9	18	27	36	45	53	63	72	81
49	69020	69108	69197	69285	69373	69461	69548	69636	69723	69810	9	18	26	35	44	53	62	70	79

LOGARITHMS

	0	1	2	3	4	5	6	7	8	9	Mean Differences								
											1	2	3	4	5	6	7	8	9
50	69897	69984	70070	70157	70243	70329	70415	70501	70586	70672	9	17	26	34	43	52	60	69	77
51	70757	70842	70927	71012	71096	71181	71265	71349	71433	71517	8	17	25	34	42	50	59	67	76
52	71600	71684	71767	71850	71933	72016	72099	72181	72263	72346	8	17	25	33	42	50	58	66	75
53	72428	72509	72591	72673	72754	72835	72916	72997	73078	73159	8	16	24	32	41	49	57	65	73
54	73239	73320	73400	73480	73560	73640	73719	73799	73878	73957	8	16	24	32	40	48	56	64	72
55	74036	74115	74194	74273	74351	74429	74507	74586	74663	74741	8	16	23	31	39	47	55	63	70
56	74819	74896	74974	75051	75128	75205	75282	75358	75435	75511	8	15	23	31	39	46	54	62	69
57	75587	75664	75740	75815	75891	75967	76042	76118	76193	76268	8	15	23	30	38	45	53	60	68
58	76343	76418	76492	76567	76641	76716	76790	76864	76938	77012	7	15	22	30	37	44	52	59	67
59	77085	77159	77232	77305	77379	77452	77525	77597	77670	77743	7	15	22	29	37	44	51	58	66
60	77815	77887	77960	78032	78104	78176	78247	78319	78390	78462	7	14	22	29	36	43	50	58	65
61	78533	78604	78675	78746	78817	78888	78958	79029	79099	79169	7	14	21	28	36	43	50	57	64
62	79239	79309	79379	79449	79518	79588	79657	79727	79796	79865	7	14	21	28	35	41	48	55	62
63	79934	80003	80072	80140	80209	80277	80346	80414	80482	80550	7	14	20	27	34	41	48	54	61
64	80618	80686	80754	80821	80889	80956	81023	81090	81158	81224	7	13	20	27	34	40	47	54	60
65	81291	81358	81425	81491	81558	81624	81690	81757	81823	81889	7	13	20	26	33	40	46	53	59
66	81954	82020	82086	82151	82217	82282	82347	82413	82478	82543	7	13	20	26	33	39	46	52	59
67	82607	82672	82737	82802	82866	82930	82995	83059	83123	83187	6	13	19	26	32	38	45	51	58
68	83251	83315	83378	83442	83506	83569	83632	83696	83759	83822	6	13	19	25	32	38	44	50	57
69	83885	83948	84011	84073	84136	84198	84261	84323	84386	84448	6	12	19	25	31	37	43	50	56
70	84510	84572	84634	84696	84757	84819	84880	84942	85003	85065	6	12	19	25	31	37	43	50	56
71	85126	85187	85248	85309	85370	85431	85491	85552	85612	85673	6	12	18	24	31	37	43	49	55
72	85733	85794	85854	85914	85974	86034	86094	86153	86213	86273	6	12	18	24	30	36	42	48	54
73	86332	86392	86451	86510	86570	86629	86688	86747	86806	86864	6	12	18	24	30	35	41	47	53
74	86923	86982	87040	87099	87157	87216	87274	87332	87390	87448	6	12	17	23	29	35	41	46	52
75	87506	87564	87622	87679	87737	87795	87852	87910	87967	88024	6	12	17	23	29	35	41	46	52
76	88081	88138	88195	88252	88309	88366	88423	88480	88536	88593	6	11	17	23	29	34	40	46	51
77	88649	88705	88762	88818	88874	88930	88986	89042	89098	89154	6	11	17	22	28	34	39	45	50
78	89209	89265	89321	89376	89432	89487	89542	89597	89653	89708	6	11	17	22	28	33	39	44	50
79	89763	89818	89873	89927	89982	90037	90091	90146	90200	90255	6	11	17	22	28	33	39	44	50
80	90309	90363	90417	90472	90526	90580	90634	90687	90741	90795	5	11	16	22	27	32	38	43	49
81	90848	90902	90956	91009	91062	91116	91169	91222	91275	91328	5	11	16	21	27	32	37	42	48
82	91381	91434	91487	91540	91593	91645	91698	91751	91803	91855	5	11	16	21	27	32	37	42	48
83	91908	91960	92012	92064	92117	92169	92221	92273	92324	92376	5	10	16	21	26	31	36	42	47
84	92428	92480	92531	92583	92634	92686	92737	92788	92840	92891	5	10	15	20	26	31	36	41	46
85	92942	92993	93044	93095	93146	93197	93247	93298	93349	93399	5	10	15	20	26	31	36	41	46
86	93450	93500	93551	93601	93651	93702	93752	93802	93852	93902	5	10	15	20	25	30	35	40	45
87	93952	94002	94052	94101	94151	94201	94250	94300	94349	94399	5	10	15	20	25	30	35	40	45
88	94448	94498	94547	94596	94645	94694	94743	94792	94841	94890	5	10	15	20	25	29	34	39	44
89	94939	94988	95036	95085	95134	95182	95231	95279	95328	95376	5	10	15	19	24	29	34	39	44
90	95424	95472	95521	95569	95617	95665	95713	95761	95809	95856	5	10	14	19	24	29	34	38	43
91	95904	95952	95999	96047	96095	96142	96190	96237	96284	96332	5	9	14	19	24	28	33	38	42
92	96379	96426	96473	96520	96567	96614	96661	96708	96755	96802	5	9	14	19	24	28	33	38	42
93	96848	96895	96942	96988	97035	97081	97128	97174	97220	97267	5	9	14	18	23	28	32	38	42
94	97313	97359	97405	97451	97497	97543	97589	97635	97681	97727	5	9	14	18	23	28	32	37	42
95	97772	97818	97864	97909	97955	98000	98046	98091	98137	98182	5	9	14	18	23	27	32	36	41
96	98227	98272	98318	98363	98408	98453	98498	98543	98588	98632	5	9	14	18	23	27	32	36	41
97	98677	98722	98767	98811	98856	98900	98945	98989	99034	99078	4	9	13	18	22	27	31	36	40
98	99123	99167	99211	99255	99300	99344	99388	99432	99476	99520	4	9	13	18	22	26	31	35	40
99	99564	99607	99651	99695	99739	99782	99826	99870	99913	99957	4	9	13	17	22	26	31	35	39

ANTILOGARITHMS

	0	1	2	3	4	5	6	7	8	9	Mean Differences								
											1	2	3	4	5	6	7	8	9
·00	10000	10023	10046	10069	10093	10116	10139	10162	10186	10209	2	5	7	9	12	14	16	19	21
·01	10233	10257	10280	10304	10328	10351	10375	10399	10423	10447	2	5	7	10	12	14	17	19	21
·02	10471	10495	10520	10544	10568	10593	10617	10641	10666	10691	2	5	7	10	12	15	17	20	22
·03	10715	10740	10765	10789	10814	10839	10864	10889	10914	10940	3	5	8	10	13	15	18	20	23
·04	10965	10990	11015	11041	11066	11092	11117	11143	11169	11194	3	5	8	10	13	15	18	20	23
·05	11220	11246	11272	11298	11324	11350	11376	11402	11429	11455	3	5	8	11	13	16	18	21	24
·06	11482	11508	11535	11561	11588	11614	11641	11668	11695	11722	3	5	8	11	13	16	19	21	24
·07	11749	11776	11803	11830	11858	11885	11912	11940	11967	11995	3	5	8	11	14	16	19	22	25
·08	12023	12050	12078	12106	12134	12162	12190	12218	12246	12274	3	6	8	11	14	17	20	22	25
·09	12303	12331	12359	12388	12417	12445	12474	12503	12531	12560	3	6	9	11	14	17	20	23	26
·10	12589	12618	12647	12677	12706	12735	12764	12794	12823	12853	3	6	9	12	15	18	21	24	26
·11	12882	12912	12942	12972	13002	13032	13062	13092	13122	13152	3	6	9	12	15	18	21	24	27
·12	13183	13213	13243	13274	13305	13335	13366	13397	13428	13459	3	6	9	12	15	18	21	25	28
·13	13490	13521	13552	13583	13614	13646	13677	13709	13740	13772	3	6	9	13	16	19	22	25	28
·14	13804	13836	13868	13900	13932	13964	13996	14028	14060	14093	3	6	10	13	16	19	22	26	29
·15	14125	14158	14191	14223	14256	14289	14322	14355	14388	14421	3	7	10	13	16	20	23	26	30
·16	14454	14488	14521	14555	14588	14622	14655	14689	14723	14757	3	7	10	13	17	20	24	27	30
·17	14791	14825	14859	14894	14928	14962	14997	15031	15066	15101	3	7	10	14	17	21	24	28	31
·18	15136	15171	15205	15241	15276	15311	15346	15382	15417	15453	4	7	11	14	18	21	25	28	32
·19	15488	15524	15560	15596	15631	15668	15704	15740	15776	15812	4	7	11	14	18	22	25	29	32
·20	15849	15885	15922	15959	15996	16032	16069	16106	16144	16181	4	7	11	15	18	22	26	30	33
·21	16218	16255	16293	16331	16368	16406	16444	16482	16520	16558	4	8	11	15	19	23	26	30	34
·22	16596	16634	16672	16711	16749	16788	16827	16866	16904	16943	4	8	12	15	19	23	27	31	35
·23	16982	17022	17061	17100	17140	17179	17219	17258	17298	17338	4	8	12	16	20	24	28	32	36
·24	17378	17418	17458	17498	17539	17579	17620	17660	17701	17742	4	8	12	16	20	24	28	32	36
·25	17783	17824	17865	17906	17947	17989	18030	18072	18113	18155	4	8	12	17	21	25	29	33	37
·26	18197	18239	18281	18323	18365	18408	18450	18493	18535	18578	4	8	13	17	21	25	30	34	38
·27	18621	18664	18707	18750	18793	18836	18880	18923	18967	19011	4	9	13	17	22	26	30	35	39
·28	19055	19099	19143	19187	19231	19275	19320	19364	19409	19454	4	9	13	18	22	26	31	35	40
·29	19498	19543	19588	19634	19679	19724	19770	19815	19861	19907	5	9	14	18	23	27	32	36	41
·30	19953	19999	20045	20091	20137	20184	20230	20277	20324	20370	5	9	14	19	23	28	32	37	42
·31	20417	20464	20512	20559	20606	20654	20701	20749	20797	20845	5	10	14	19	24	29	33	38	43
·32	20893	20941	20989	21038	21086	21135	21184	21232	21281	21330	5	10	15	19	24	29	34	39	44
·33	21380	21429	21478	21528	21577	21627	21677	21727	21777	21827	5	10	15	20	25	30	35	40	45
·34	21878	21928	21979	22029	22080	22131	22182	22233	22284	22336	5	10	15	20	25	31	36	41	46
·35	22387	22439	22491	22542	22594	22646	22699	22751	22803	22856	5	10	16	21	26	31	37	42	47
·36	22909	22961	23014	23067	23121	23174	23227	23281	23336	23388	5	11	16	21	27	32	37	43	48
·37	23442	23496	23550	23605	23659	23714	23768	23823	23878	23933	5	11	16	22	27	33	38	44	49
·38	23988	24044	24099	24155	24210	24266	24322	24378	24434	24491	6	11	17	22	28	34	39	45	50
·39	24547	24604	24660	24717	24774	24831	24889	24946	25003	25061	6	11	17	23	29	34	40	46	51
·40	25119	25177	25236	25293	25351	25410	25468	25527	25586	25645	6	12	18	23	29	35	41	47	53
·41	25704	25763	25823	25882	25942	26002	26062	26122	26182	26242	6	12	18	24	30	36	42	48	54
·42	26303	26363	26424	26485	26546	26607	26669	26730	26792	26853	6	12	18	24	31	37	43	49	55
·43	26915	26977	27040	27102	27164	27227	27290	27353	27416	27479	6	13	19	25	31	38	44	50	56
·44	27542	27606	27669	27733	27797	27861	27925	27990	28054	28119	6	13	19	26	32	39	45	51	58
·45	28184	28249	28314	28379	28445	28510	28576	28642	28708	28774	7	13	20	26	33	39	46	52	59
·46	28840	28907	28973	29040	29107	29174	29242	29309	29376	29444	7	13	20	27	34	40	47	54	60
·47	29512	29580	29648	29717	29785	29854	29923	29992	30061	30130	7	14	21	28	34	41	48	55	62
·48	30200	30259	30339	30409	30479	30549	30620	30690	30761	30832	7	14	21	28	35	42	49	56	63
·49	30903	30974	31046	31117	31189	31261	31333	31405	31477	31550	7	14	22	29	36	43	50	58	65

ANTILOGARITHMS

	0	1	2	3	4	5	6	7	8	9	Mean Differences								
											1	2	3	4	5	6	7	8	9
·50	31623	31696	31769	31842	31916	31989	32063	32137	32211	32285	7	15	22	29	37	44	52	59	66
·51	32359	32434	32509	32584	32659	32735	32809	32885	32961	33037	8	15	23	30	38	45	53	60	68
·52	33113	33189	33266	33343	33420	33497	33574	33651	33729	33806	8	15	23	31	39	46	54	62	69
·53	33884	33963	34041	34119	34198	34277	34356	34435	34514	34594	8	16	24	32	40	47	55	63	71
·54	34674	34754	34834	34914	34995	35075	35156	35237	35318	35400	8	16	24	32	40	48	56	65	73
·55	35481	35563	35645	35727	35810	35892	35975	36058	36141	36224	8	16	25	33	41	50	58	66	74
·56	36308	36392	36475	36559	36644	36728	36813	36898	36983	37068	8	17	25	34	42	51	59	68	76
·57	37154	37239	37325	37411	37497	37584	37670	37757	37844	37931	9	17	26	35	43	52	61	69	78
·58	38019	38107	38194	38282	38371	38459	38548	38637	38726	38815	9	18	27	35	44	53	62	71	80
·59	38905	3899	39084	39174	39264	39355	39446	39537	39628	39719	9	18	27	36	45	54	63	72	82
·60	39811	39902	39994	40087	40179	40272	40365	40458	40551	40644	9	19	28	37	46	56	65	74	83
·61	40738	40832	40926	41020	41115	41210	41305	41400	41495	41591	9	19	28	38	47	57	66	76	85
·62	41687	41783	41879	41976	42073	42170	42267	42364	42462	42560	10	19	29	39	49	58	68	78	87
·63	42658	42756	42855	42954	43053	43152	43251	43351	43451	43551	10	20	30	40	50	60	70	80	89
·64	43652	43752	43853	43954	44055	44157	44259	44361	44463	44566	10	20	30	41	51	61	71	81	91
·65	44668	44771	44875	44978	45082	45186	45290	45394	45499	45604	10	21	31	42	52	62	73	83	94
·66	45709	45814	45920	46026	46132	46238	46345	46452	46559	46666	11	21	32	43	53	64	75	85	96
·67	46774	46881	46989	47098	47206	47315	47424	47534	47643	47753	11	22	33	44	54	65	76	87	98
·68	47863	47973	48084	48195	48306	48417	48529	48641	48753	48865	11	22	33	45	56	67	78	89	100
·69	48978	49091	49204	49317	49431	49545	49659	49774	49888	50003	11	23	34	46	57	68	80	91	103
·70	50119	50234	50350	50466	50582	50699	50816	50933	51050	51168	12	23	35	47	58	70	82	93	105
·71	51286	51404	51523	51642	51761	51880	52000	52119	52240	52360	12	24	36	48	60	72	84	96	108
·72	52481	52602	52723	52845	52966	53088	53211	53333	53456	53580	12	24	37	49	61	73	85	98	110
·73	53703	53827	53951	54075	54200	54325	54450	54576	54702	54828	13	25	38	50	63	75	88	100	113
·74	54954	55081	55208	55336	55463	55590	55719	55847	55976	56105	13	26	38	51	64	77	90	102	115
·75	56234	56364	56494	56624	56754	56885	57016	57148	57280	57412	13	26	39	52	66	79	92	105	118
·76	57544	57677	57810	57943	58076	58210	58345	58479	58614	58749	13	27	40	54	67	80	94	107	121
·77	58884	59020	59156	59293	59429	59566	59704	59841	59979	60117	14	27	41	55	69	82	96	110	123
·78	60256	60395	60534	60674	60814	60954	61094	61235	61376	61518	14	28	42	56	70	84	98	112	126
·79	61659	61802	61944	62087	62230	62373	62517	62661	62806	62951	14	29	43	58	72	86	101	115	130
·80	63096	63241	63387	63533	63680	63826	63973	64121	64269	64417	15	29	44	59	74	88	103	118	132
·81	64565	64714	64863	65013	65163	65313	65464	65615	65766	65917	15	30	45	60	75	90	105	120	135
·82	66069	66222	66374	66527	66681	66834	66988	67143	67298	67453	15	31	46	62	77	92	108	123	139
·83	67608	67764	67920	68077	68234	68391	68549	68707	68865	69024	16	32	47	63	79	95	110	126	142
·84	69183	69343	69503	69663	69823	69984	70146	70307	70469	70632	16	32	48	64	81	97	113	129	145
·85	70795	70958	71121	71285	71450	71614	71779	71945	72111	72277	17	33	50	66	83	99	116	132	149
·86	72444	72611	72778	72946	73114	73282	73451	73621	73790	73961	17	34	51	68	85	101	118	135	152
·87	74131	74302	74473	74645	74817	74989	75162	75336	75509	75683	17	35	52	69	87	104	121	138	156
·88	75858	76033	76208	76384	76560	76736	76913	77090	77268	77446	18	35	53	71	89	107	125	142	159
·89	77625	77804	77983	78163	78343	78524	78705	78886	79068	79250	18	36	54	72	91	109	127	145	163
·90	79433	79616	79799	79983	80168	80353	80538	80724	80910	81096	19	37	56	74	93	111	130	148	167
·91	81283	81470	81658	81846	82035	82224	82414	82604	82794	82985	19	38	57	76	95	113	132	151	170
·92	83176	83368	83560	83753	83946	84140	84333	84528	84723	84918	19	39	58	78	97	116	136	155	175
·93	85114	85310	85507	85704	85901	86099	86298	86497	86696	86896	20	40	60	79	99	119	139	158	178
·94	87096	87297	87498	87700	87902	88105	88308	88512	88716	88920	20	41	61	81	102	122	142	162	183
·95	89125	89331	89536	89743	89950	90157	90365	90573	90782	90991	21	42	62	83	104	125	146	166	187
·96	91201	91411	91622	91833	92045	92257	92470	92683	92897	93111	21	42	64	85	106	127	149	170	191
·97	93325	93541	93756	93972	94189	94406	94624	94842	95060	95280	22	43	65	87	109	130	152	174	195
·98	95499	95719	95940	96161	96383	96605	96828	97051	97275	97499	22	44	67	89	111	133	155	178	200
·99	97724	97949	98175	98401	98628	98855	99083	99312	99541	99770	23	46	68	91	114	137	160	182	205

NATURAL SINES

Degrees	0′	6′	12′	18′	24′	30′	36′	42′	48′	54′	Mean Differences				
											1	2	3	4	5
0	·00000	00175	00349	00524	00698	00873	01047	01222	01396	01571	29	58	87	117	145
1	·01745	01920	02094	02269	02443	02618	02792	02967	03141	03316	29	58	87	117	145
2	·03490	03664	03839	04013	04188	04362	04536	04711	04885	05059	29	58	87	116	145
3	·05234	05408	05582	05756	05931	06105	06279	06453	06627	06802	29	58	87	116	145
4	·06976	07150	07324	07498	07672	07846	08020	08194	08368	08542	29	58	87	116	145
5	·08716	08889	09063	09237	09411	09585	09758	09932	10106	10279	29	58	87	116	145
6	·10453	10626	10800	10973	11147	11320	11494	11667	11840	12014	29	58	87	116	145
7	·12187	12360	12533	12706	12880	13053	13226	13399	13572	13744	29	58	87	115	144
8	·13917	14090	14263	14436	14608	14781	14954	15126	15299	15471	29	58	86	115	144
9	·15643	15816	15988	16160	16333	16505	16677	16849	17021	17193	29	57	86	115	144
10	·17365	17537	17708	17880	18052	18224	18395	18567	18738	18910	29	57	86	114	143
11	·19081	19252	19423	19595	19766	19937	20108	20279	20450	20620	29	57	86	114	143
12	·20791	20962	21132	21303	21474	21644	21814	21985	22155	22325	28	57	85	114	142
13	·22495	22665	22835	23005	23175	23345	23514	23684	23853	24023	28	57	85	113	141
14	·24192	24362	24531	24700	24869	25038	25207	25376	25545	25713	28	56	85	113	141
15	·25882	26050	26219	26387	26556	26724	26892	27060	27228	27396	28	56	84	112	140
16	·27564	27731	27899	28067	28234	28402	28569	28736	28903	29070	28	56	84	112	139
17	·29237	29404	29571	29737	29904	30071	30237	30403	30570	30736	28	56	83	111	139
18	·30902	31068	31233	31399	31565	31730	31896	32061	32227	32392	28	55	83	110	138
19	·32557	32722	32887	33051	33216	33381	33545	33710	33874	34038	27	55	82	110	137
20	·34202	34366	34530	34694	34857	35021	35184	35347	35511	35674	27	55	82	109	136
21	·35837	36000	36162	36325	36488	36650	36812	36975	37137	37299	27	54	81	108	135
22	·37461	37622	37784	37946	38107	38268	38430	38591	38752	38912	27	54	81	107	134
23	·39073	39234	39394	39555	39715	39875	40035	40195	40355	40514	27	53	80	107	133
24	·40674	40833	40992	41151	41310	41469	41628	41787	41945	42104	27	53	79	106	132
25	·42262	42420	42578	42736	42894	43051	43209	43366	43523	43680	26	53	79	105	131
26	·43837	43994	44151	44307	44464	44620	44776	44932	45088	45243	26	52	78	104	130
27	·45399	45554	45710	45865	46020	46175	46330	46484	46639	46793	26	52	77	103	129
28	·46947	47101	47255	47409	47562	47716	47869	48022	48175	48328	26	51	77	102	128
29	·48481	48634	48786	48938	49090	49242	49394	49546	49697	49849	25	51	76	101	126
30	·50000	50151	50302	50453	50603	50754	50904	51054	51204	51354	25	50	75	100	125
31	·51504	51653	51803	51952	52101	52250	52399	52547	52696	52844	25	50	74	99	124
32	·52992	53140	53288	53435	53583	53730	53877	54024	54171	54317	25	49	74	98	123
33	·54464	54610	54756	54902	55048	55194	55339	55484	55630	55775	24	49	73	97	121
34	·55919	56064	56208	56353	56497	56641	56784	56928	57071	57215	24	48	72	96	120
35	·57358	57501	57643	57786	57928	58070	58212	58354	58496	58637	24	48	71	95	118
36	·58778	58920	59061	59201	59342	59482	59622	59763	59902	60042	23	47	70	94	117
37	·60181	60321	60460	60599	60738	60876	61015	61153	61291	61429	23	46	69	92	115
38	·61566	61704	61841	61978	62115	62251	62388	62524	62660	62796	23	46	68	91	114
39	·62932	63068	63203	63338	63473	63608	63742	63877	64011	64145	23	45	67	90	112
40	·64279	64412	64546	64679	64812	64945	65077	65210	65342	65474	22	44	66	88	111
41	·65606	65738	65869	66000	66131	66262	66393	66523	66653	66783	22	44	65	87	109
42	·66913	67043	67172	67301	67430	67559	67688	67816	67944	68072	22	43	64	86	107
43	·68200	68327	68455	68582	68709	68835	68962	69088	69214	69340	21	42	63	84	106
44	·69466	69591	69717	69842	69966	70091	70215	70339	70463	70587	21	41	62	83	104

NATURAL SINES

Degrees	0′	6′	12′	18′	24′	30′	36′	42′	48′	54′	Mean Differences				
											1	2	3	4	5
45	·70711	70834	70957	71080	71203	71325	71447	71569	71691	71813	20	41	61	82	102
46	·71934	72055	72176	72297	72417	72537	72657	72777	72897	73016	20	40	60	80	100
47	·73135	73254	73373	73491	73610	73728	73846	73963	74080	74198	20	39	59	79	98
48	·74314	74431	74548	74664	74780	74896	75011	75126	75241	75356	20	39	58	77	96
49	·75471	75585	75700	75813	75927	76041	76154	76267	76380	76492	19	38	57	76	94
50	·76604	76717	76828	76940	77051	77162	77273	77384	77494	77605	19	37	56	74	93
51	·77715	77824	77934	78043	78152	78261	78369	78478	78586	78694	18	36	54	72	91
52	·78801	78908	79015	79122	79229	79335	79441	79547	79653	79758	18	35	53	71	89
53	·79864	79968	80073	80178	80282	80386	80489	80593	80696	80799	17	35	52	69	87
54	·80902	81004	81106	81208	81310	81412	81513	81614	81714	81815	17	34	51	68	84
55	·81915	82015	82115	82214	82314	82413	82511	82610	82708	82806	16	33	49	66	82
56	·82904	83001	83098	83195	83292	83389	83485	83581	83676	83772	16	32	48	64	80
57	·83867	83962	84057	84151	84245	84339	84433	84526	84619	84712	16	31	47	62	78
58	·84805	84897	84987	85081	85173	85264	85355	85446	85536	85627	15	31	46	61	76
59	·85717	85806	85896	85985	86074	86163	86251	86340	86427	86515	15	30	44	59	74
60	·86603	86690	86777	86863	86949	87036	87121	87207	87292	87377	14	29	43	57	72
61	·87462	87546	87631	87715	87798	87882	87965	88048	88130	88213	14	28	42	56	69
62	·88295	88377	88458	88539	88620	88701	88782	88862	88942	89021	14	27	40	54	67
63	·89101	89180	89259	89337	89415	89493	89571	89649	89726	89803	13	26	39	52	65
64	·89879	89956	90032	90108	90183	90259	90334	90408	90483	90557	13	25	38	50	63
65	·90631	90704	90778	90851	90924	90996	91068	91140	91212	91283	12	24	36	48	60
66	·91355	91425	91496	91566	91636	91706	91775	91845	91914	91982	12	23	35	46	58
67	·92050	92119	92186	92254	92321	92388	92455	92521	92587	92653	11	22	33	45	56
68	·92718	92784	92849	92913	92978	93042	93106	93169	93232	93295	11	21	32	43	54
69	·93358	93420	93483	93544	93606	93667	93728	93789	93849	93909	10	20	31	41	51
70	·93969	94029	94088	94147	94206	94264	94322	94380	94438	94495	10	19	29	39	49
71	·94552	94609	94665	94721	94777	94832	94888	94943	94997	95052	9	18	28	37	46
72	·95106	95159	95213	95266	95319	95372	95424	95476	95528	95579	9	17	26	35	44
73	·95630	95681	95732	95782	95832	95882	95931	95981	96029	96078	8	17	25	33	41
74	·96126	96174	96222	96269	96316	96363	96410	96456	96502	96547	8	16	23	31	39
75	·96593	96638	96682	96727	96771	96815	96858	96902	96945	96987	7	16	22	29	36
76	·97030	97072	97113	97155	97196	97237	97278	97318	97358	97398	7	14	20	27	34
77	·97437	97476	97515	97553	97592	97630	97667	97705	97742	97778	6	13	19	25	32
78	·97815	97851	97887	97922	97958	97992	98027	98061	98096	98129	6	12	17	23	29
79	·98163	98196	98229	98261	98294	98325	98357	98388	98420	98450	5	11	16	21	27
80	·98481	98511	98541	98570	98600	98629	98657	98686	98714	98741	5	10	14	19	24
81	·98769	98796	98823	98849	98876	98902	98927	98953	98978	99002	4	9	13	17	22
82	·99027	99051	99075	99098	99122	99144	99167	99189	99211	99233	4	8	11	15	19
83	·99255	99276	99297	99317	99337	99357	99377	99396	99415	99434	3	7	10	13	16
84	·99452	99470	99488	99506	99523	99540	99556	99572	99588	99604	3	6	8	11	14
85	·99619	99635	99649	99664	99678	99692	99705	99719	99731	99744	2	5	7	9	11
86	·99756	99768	99780	99792	99803	99813	99824	99834	99844	99854	2	4	5	7	9
87	·99863	99872	99881	99889	99897	99905	99912	99919	99926	99933	1	3	4	5	6
88	·99939	99945	99951	99956	99961	99966	99970	99974	99978	99982	1	2	2	3	4
89	·99985	99988	99990	99993	99995	99996	99998	99999	99999	99999	0	1	1	1	1
90	1·00000														

NATURAL COSINES

Degrees	0′	6′	12′	18′	24′	30′	36′	42′	48′	54′	Mean Differences				
											1	2	3	4	5
0	1·00000	99999	99999	99999	99998	99996	99995	99993	99990	99988	0	1	1	1	1
1	·99985	99982	99978	99974	99970	99966	99961	99956	99951	99945	1	2	2	3	4
2	·99939	99933	99926	99919	99912	99905	99897	99889	99881	99872	1	3	4	5	6
3	·99863	99854	99844	99834	99824	99813	99803	99792	99780	99768	2	4	5	7	9
4	·99756	99744	99731	99719	99705	99692	99678	99664	99649	99635	2	5	7	9	11
5	·99619	99604	99588	99572	99556	99540	99523	99506	99488	99470	3	6	8	11	14
6	·99452	99434	99415	99396	99377	99357	99337	99317	99297	99276	3	7	10	13	16
7	·99255	99233	99211	99189	99167	99144	99122	99098	99075	99051	4	8	11	15	19
8	·99027	99002	98978	98953	98927	98902	98876	98849	98823	98796	4	9	13	17	22
9	·98769	98741	98714	98686	98657	98629	98600	98570	98541	98511	5	10	14	19	24
10	·98481	98450	98420	98388	98357	98325	98294	98261	98229	98196	5	11	16	21	27
11	·98163	98129	98096	98061	98027	97992	97958	97922	97887	97851	6	12	17	23	29
12	·97815	97778	97742	97705	97667	97630	97592	97553	97515	97476	6	13	19	25	32
13	·97437	97398	97358	97318	97278	97237	97196	97155	97113	97072	7	14	20	27	34
14	·97030	96987	96945	96902	96858	96815	96771	96727	96682	96638	7	16	22	29	36
15	·96593	96547	96502	96456	96410	96363	96316	96269	96222	96174	8	16	23	31	39
16	·96126	96078	96029	95981	95931	95882	95832	95782	95732	95681	8	17	25	33	41
17	·95630	95579	95528	95476	95424	95372	95319	95266	95213	95159	9	17	26	35	44
18	·95106	95052	94997	94943	94888	94832	94777	94721	94665	94609	9	18	28	37	46
19	·94552	94495	94438	94380	94322	94264	94206	94147	94088	94029	10	19	29	39	49
20	·93969	93909	93849	93789	93728	93667	93606	93544	93483	93420	10	20	31	41	51
21	·93358	93295	93232	93169	93106	93042	92978	92913	92849	92784	10	21	32	43	54
22	·92718	92653	92587	92521	92455	92388	92321	92254	92186	92119	11	22	33	45	56
23	·92050	91982	91914	91845	91775	91706	91636	91566	91496	91425	11	23	35	46	58
24	·91355	91283	91212	91140	91068	90996	90924	90851	90778	90704	12	24	36	48	60
25	·90631	90557	90483	90408	90334	90259	90183	90108	90032	89956	13	25	38	50	63
26	·89879	89803	89726	89649	89571	89493	89415	89337	89259	89180	13	26	39	52	65
27	·89101	89021	88942	88862	88782	88701	88620	88539	88458	88377	13	27	40	54	67
28	·88295	88213	88130	88048	87965	87882	87798	87715	87631	87546	14	28	42	56	69
29	·87462	87377	87292	87207	87121	87036	86949	86863	86777	86690	14	29	43	57	72
30	·86603	86515	86427	86340	86251	86163	86074	85985	85896	85806	15	30	44	59	74
31	·85717	85627	85536	85446	85355	85264	85173	85081	84989	84897	16	30	46	61	76
32	·84805	84712	84619	84526	84433	84339	84245	84151	84057	83962	16	31	47	62	78
33	·83867	83772	83676	83581	83485	83389	83292	83195	83098	83001	16	32	48	64	80
34	·82904	82806	82708	82610	82511	82413	82314	82214	82115	82015	16	33	49	66	82
35	·81915	81815	81714	81614	81513	81412	81310	81208	81106	81004	17	34	51	68	84
36	·80902	80799	80696	80593	80489	80386	80282	80178	80073	79968	17	35	52	69	87
37	·79864	79758	79653	79547	79441	79335	79229	79122	79015	78908	18	35	53	71	89
38	·78801	78693	78586	78478	78369	78261	78152	78043	77934	77824	18	36	54	72	91
39	·77715	77605	77494	77384	77273	77162	77051	76940	76828	76717	18	37	56	74	93
40	·76604	76492	76380	76267	76154	76041	75927	75813	75700	75585	19	38	57	76	94
41	·75471	75356	75241	75126	75011	74896	74780	74664	74548	74431	19	39	58	77	96
42	·74314	74198	74080	73963	73846	73728	73610	73491	73373	73254	20	39	59	79	98
43	·73135	73016	72897	72777	72657	72537	72417	72297	72176	72055	20	40	60	80	100
44	·71934	71813	71691	71569	71447	71325	71203	71080	70957	70834	20	41	61	82	102

NATURAL COSINES

Degrees	0′	6′	12′	18′	24′	30′	36′	42′	48′	54′	Mean Differences				
											1	2	3	4	5
45	·70711	70587	70463	70339	70215	70091	69966	69842	69717	69591	21	41	62	83	104
46	·69466	69340	69214	69088	68962	68835	68709	68582	68455	68327	21	42	63	84	106
47	·68200	68072	67944	67816	67688	67559	67430	67301	67172	67043	21	43	64	86	107
48	·66913	66783	66653	66523	66393	66262	66131	66000	65869	65738	22	44	65	87	109
49	·65606	65474	65342	65210	65077	64945	64812	64679	64546	64412	22	44	66	88	111
50	·64279	64145	64011	63877	63742	63608	63473	63338	63203	63068	22	45	67	90	112
51	·62932	62796	62660	62524	62388	62251	62115	61978	61841	61704	23	45	68	91	114
52	·61566	61429	61291	61153	61015	60876	60738	60599	60460	60321	23	46	69	92	115
53	·60181	60042	59902	59763	59622	59482	59342	59201	59061	58920	23	47	70	94	117
54	·58779	58637	58496	58354	58212	58070	57928	57786	57643	57501	24	47	71	95	118
55	·57358	57215	57071	56928	56784	56641	56497	56353	56208	56064	24	48	72	96	120
56	·55919	55775	55630	55484	55339	55194	55048	54902	54756	54610	24	49	73	97	121
57	·54464	54317	54171	54024	53877	53730	53583	53435	53288	53140	24	49	74	98	123
58	·52992	52844	52696	52547	52399	52250	52101	51952	51803	51653	25	50	74	99	124
59	·51504	51354	51204	51054	50904	50754	50603	50453	50302	50151	25	50	75	100	125
60	·50000	49849	49697	49546	49394	49242	49090	48938	48786	48634	25	51	76	101	126
61	·48481	48328	48175	48022	47869	47716	47562	47409	47255	47101	26	51	77	102	128
62	·46947	46793	46639	46484	46330	46175	46020	45865	45710	45554	26	52	77	103	129
63	·45399	45243	45088	44932	44776	44620	44464	44307	44151	43994	26	52	78	104	130
64	·43837	43680	43523	43366	43209	43051	42894	42736	42578	42420	26	53	79	105	131
65	·42262	42104	41945	41787	41628	41469	41310	41151	40992	40833	26	53	79	106	132
66	·40674	40514	40355	40195	40035	39875	39715	39555	39394	39234	27	53	80	107	133
67	·39073	38912	38752	38591	38430	38268	38107	37946	37784	37622	27	54	81	107	134
68	·37461	37299	37137	36975	36812	36650	36488	36325	36162	36000	27	54	81	108	135
69	·35837	35674	35511	35347	35184	35021	34857	34694	34530	34366	27	54	82	109	136
70	·34202	34038	33874	33710	33545	33381	33216	33051	32887	33722	27	55	82	110	137
71	·32557	32392	32227	32061	31896	31730	31565	31399	31233	31068	28	55	83	110	138
72	·30902	30736	30570	30403	30237	30071	29904	29737	29571	29404	28	56	83	111	139
73	·29237	29070	28903	28736	28569	28402	28234	28067	27899	27731	28	56	84	112	139
74	·27564	27396	27228	27060	26892	26724	26556	26387	26219	26050	28	56	84	112	140
75	·25882	25713	25545	25376	25207	25038	24869	24700	24531	24362	28	56	85	113	141
76	·24192	24023	23853	23684	23514	23345	23175	23005	22835	22665	28	57	85	113	141
77	·22495	22325	22155	21985	21814	21644	21474	21303	21132	20962	28	57	85	114	142
78	·20791	20620	20450	20279	20108	19937	19766	19595	19423	19252	29	57	86	114	143
79	·19081	18910	18738	18567	18395	18224	18052	17880	17708	17537	29	57	86	114	143
80	·17365	17193	17021	16849	16677	16505	16333	16160	15988	15816	29	57	86	115	144
81	·15643	15471	15299	15126	14954	14781	14608	14436	14263	14090	29	58	86	115	144
82	·13917	13744	13572	13399	13226	13053	12880	12706	12533	12360	29	58	87	115	144
83	·12187	12014	11840	11667	11494	11320	11147	10973	10800	10626	29	58	87	116	145
84	·10453	10279	10106	09932	09758	09585	09411	09237	09063	08889	29	58	87	116	145
85	·08716	08542	08368	08194	08020	07846	07672	07498	07324	07150	29	58	87	116	145
86	·06976	06802	06627	06453	06279	06105	05931	05756	05582	05408	29	58	87	116	145
87	·05234	05059	04885	04711	04536	04362	04188	04013	03839	03664	29	58	87	116	145
88	·03490	03316	03141	02967	02792	02618	02443	02269	02094	01920	29	58	87	117	145
89	·01745	01571	01396	01222	01047	00873	00698	00524	00349	00175	29	58	87	117	145
90	·00000														

NATURAL TANGENTS

Degrees	0′	6′	12′	18′	24′	30′	36′	42′	48′	54′	Mean Differences				
											1	2	3	4	5
0	·00000	00175	00349	00524	00698	00873	01047	01222	01396	01571	29	58	87	116	146
1	·01746	01920	02095	02269	02444	02619	02793	02968	03143	03317	29	58	87	116	146
2	·03492	03667	03842	04016	04191	04366	04541	04716	04891	05066	29	58	87	117	146
3	·05241	05416	05591	05766	05941	06116	06291	06467	06642	06817	29	58	88	117	146
4	·06993	07168	07344	07519	07695	07870	08046	08221	08397	08573	29	59	88	117	146
5	·08749	08925	09101	09277	09453	09629	09805	09981	10158	10334	29	59	88	117	147
6	·10510	10687	10863	11040	11217	11394	11570	11747	11924	12101	29	59	88	118	147
7	·12278	12456	12633	12810	12988	13165	13343	13521	13698	13876	30	59	89	118	148
8	·14054	14232	14410	14588	14767	14945	15124	15302	15481	15660	30	59	89	119	149
9	·15838	16017	16196	16376	16555	16734	16914	17093	17273	17453	30	60	90	120	150
10	·17633	17813	17993	18173	18353	18534	18714	18895	19076	19257	30	60	90	120	150
11	·19438	19619	19801	19982	20164	20345	20527	20709	20891	21073	30	60	91	121	152
12	·21256	21438	21621	21804	21986	22169	22353	22536	22719	22903	30	61	92	122	153
13	·23087	23271	23455	23639	23823	24008	24193	24377	24562	24747	31	61	93	124	155
14	·24933	25118	25304	25490	25676	25862	26048	26235	26421	26608	31	62	93	124	155
15	·26795	26982	27169	27357	27545	27732	27920	28109	28297	28486	31	63	94	125	157
16	·28675	28864	29053	29242	29432	29621	29811	30001	30192	30382	32	63	95	127	158
17	·30573	30764	30955	31147	31338	31530	31722	31914	32106	32299	32	64	96	128	160
18	·32492	32685	32878	33072	33266	33460	33654	33848	34043	34238	32	65	97	129	162
19	·34433	34628	34824	35019	35216	35412	35608	35805	36002	36199	33	66	98	131	164
20	·36397	36595	36793	36991	37190	37388	37588	37787	37986	38186	33	66	99	133	166
21	·38386	38587	38787	38988	39190	39391	39593	39795	39997	40200	34	67	101	134	168
22	·40403	40606	40809	41013	41217	41421	41626	41831	42036	42242	34	68	102	136	170
23	·42447	42654	42860	43067	43274	43481	43689	43897	44105	44314	34	69	104	138	173
24	·44523	44732	44942	45152	45362	45573	45784	45995	46206	46418	35	70	105	141	176
25	·46631	46843	47056	47270	47483	47698	47912	48127	48342	48557	36	71	107	143	179
26	·48773	48989	49206	49423	49640	49858	50076	50295	50514	50733	36	73	109	145	182
27	·50953	51173	51393	51614	51835	52057	52279	52501	52724	52947	37	74	111	148	185
28	·53171	53395	53620	53844	54070	54296	54522	54748	54975	55203	38	75	113	151	188
29	·55431	55659	55888	56117	56347	56577	56808	57039	57271	57503	38	77	115	154	192
30	·57735	57968	58201	58435	58670	58905	59140	59376	59612	59849	39	78	118	157	196
31	·60086	60324	60562	60801	61040	61280	61520	61761	62003	62245	40	79	120	160	200
32	·62487	62730	62973	63217	63462	63707	63953	64199	64446	64693	41	82	123	164	205
33	·64941	65189	65438	65688	65938	66189	66440	66692	66944	67197	42	84	126	167	209
34	·67451	67705	67960	68215	68471	68728	68985	69243	69502	69761	43	86	129	171	214
35	·70021	70281	70542	70804	71066	71329	71593	71857	72122	72388	44	88	132	176	219
36	·72654	72921	73189	73457	73726	73996	74267	74538	74810	75082	45	90	135	180	225
37	·75355	75629	75904	76180	76456	76733	77010	77289	77568	77848	46	92	139	185	231
38	·78129	78410	78692	78975	79259	79544	79829	80115	80402	80690	47	95	142	190	237
39	·80978	81268	81558	81849	82141	82434	82727	83022	83317	83613	49	98	147	195	244
40	·83910	84208	84507	84806	85107	85408	85710	86014	86318	86623	50	100	151	201	252
41	·86929	87236	87543	87852	88162	88473	88784	89097	89410	89725	52	103	155	207	259
42	·90040	90357	90674	90993	91313	91633	91955	92277	92601	92926	53	107	160	214	268
43	·93252	93578	93906	94235	94565	94896	95229	95562	95897	96232	55	111	165	221	276
44	·96569	96907	97246	97586	97927	98270	98613	98950	99304	99652	57	114	171	229	286

NATURAL TANGENTS

Degrees	0′	6′	12′	18′	24′	30′	36′	42′	48′	54′	Mean Differences				
											1	2	3	4	5
45	1·00000	00350	00701	01053	01406	01761	02117	02474	02832	03192	58	118	177	237	296
46	1·03553	03915	04279	04644	05010	05378	05747	06117	06489	06862	61	123	184	245	307
47	1·07237	07613	07990	08369	08749	09131	09514	09899	10285	10672	63	127	191	255	319
48	1·11061	11452	11844	12238	12633	13029	13428	13828	14229	14632	66	132	199	265	331
49	1·15037	15443	15851	16261	16672	17085	17500	17916	18334	18754	69	138	207	276	344
50	1·19175	19599	20024	20451	20879	21310	21742	22176	22612	23050	72	143	216	288	359
51	1·23490	23931	24375	24820	25268	25717	26169	26622	27077	27535	75	150	225	300	375
52	1·27994	28456	28919	29385	29853	30323	30795	31269	31745	32224	78	157	235	314	392
53	1·32704	33187	33673	34160	34650	35142	35637	36134	36633	37134	82	164	247	329	411
54	1·37638	38145	38653	39165	39679	40195	40714	41235	41759	42286	86	172	259	345	431
55	1·42815	43347	43881	44418	44958	45501	46046	46595	47146	47700	91	181	272	362	453
56	1·48256	48816	49378	49944	50512	51084	51658	52235	52816	53400	95	191	286	382	477
57	1·53987	54576	55170	55767	56366	56969	57575	58184	58797	59414	100	201	302	403	504
58	1·60033	60657	61283	61914	62548	63185	63826	64471	65120	65772	106	213	319	426	533
59	1·66428	67088	67752	68419	69091	69766	70446	71129	71817	72509	113	226	339	452	564
60	1·73205	73905	74610	75319	76032	76749	77471	78198	78929	79665	120	240	360	481	600
61	1·80405	81150	81900	82654	83413	84177	84946	85720	86500	87283	128	255	383	511	639
62	1·88073	88867	89667	90472	91282	92098	92920	93746	94579	95417	136	273	409	546	683
63	1·96261	97111	97967	98828	99695	2·00569	2·01449	2·02335	2·03227	2·04125	146	292	438	584	731
64	2·05030	05942	06860	07785	08716	09654	10600	11552	12511	13477	157	314	471	629	786
65	2·14451	15432	16420	17416	18419	19430	20449	21475	22510	23553	169	338	508	677	846
66	2·24604	25663	26730	27806	28891	29984	31086	32197	33317	34447	183	366	549	732	915
67	2·35585	36733	37891	39058	40235	41421	42618	43825	45043	46270	199	397	596	795	994
68	2·47509	48758	50018	51289	52571	53865	55170	56487	57815	59156					
69	2·60509	61874	63252	64642	66046	67462	68892	70335	71792	73263					
70	2·74748	76247	77761	79289	80833	82391	83965	85556	87161	88783					
71	2·90421	92076	93748	95437	97144	98868	3·00611	3·02372	3·04152	3·05950					
72	3·07768	09606	11464	13341	15240	17159	19100	21063	23048	25055					
73	3·27085	29139	31216	33317	35443	37594	39771	41973	44202	46458					
74	3·48741	51053	53393	55761	58160	60588	63048	65538	68061	70616					
75	3·73205	75828	78485	81177	83906	86671	89474	92316	95196	98117					
76	4·01078	04081	07127	10216	13350	16530	19756	23030	26352	29724					
77	4·33148	36623	40152	43735	47374	51071	54826	58641	62518	66458					
78	4·70463	74534	78673	82882	87162	91516	95945	5·00451	5·05037	5·09704					
79	5·14455	19293	24218	29235	34345	39552	44857	50264	55777	61397					
80	5·67128	72974	78938	85024	91236	97576	6·04051	6·10664	6·17419	6·24321					
81	6·31375	38587	45961	53503	61220	69116	77199	85475	93952	7·02637					
82	7·11537	20661	30018	39616	49465	59575	69957	80622	91582	8·02848					
83	8·14435	26356	38625	51259	64275	77689	91520	9·05789	9·20516	9·35724					
84	9·51436	9·6768	9·8448	10·019	10·199	10·385	10·579	10·780	10·988	11·205					
85	11·4301	11·664	11·909	12·163	12·429	12·706	12·996	13·300	13·617	13·951					
86	14·3007	14·669	15·056	15·464	15·895	16·350	16·832	17·343	17·886	18·464					
87	19·0811	19·740	20·446	21·205	22·022	22·904	23·859	24·898	26·031	27·271					
88	28·6363	30·145	31·821	33·694	35·801	38·188	40·917	44·066	47·740	52·081					
89	57·2900	63·657	71·615	81·847	95·489	114·589	143·237	190·984	286·48	572·96					
90	∞														

NATURAL COSECANTS

Degrees	0′	6′	12′	18′	24′	30′	36′	42′	48′	54′	Mean Differences				
											1	2	3	4	5
0	∞	572·96	286·48	190·99	143·24	144·59	95·495	81·853	71·622	63·665					
1	57·299	52·090	47·750	44·077	40·930	38·202	35·815	33·708	31·836	30·161					
2	28·654	27·290	26·050	24·918	23·880	22·926	22·044	21·229	20·471	19·766					
3	19·107	18·492	17·914	17·372	16·862	16·380	15·926	15·496	15·089	14·703					
4	14·336	13·987	13·654	13·337	13·035	12·745	12·469	12·204	11·951	11·707					
5	11·474	11·2493	11·0336	10·8260	10·6261	10·4334	10·2477	10·0685	9·8955	9·7283					
6	9·5668	9·4105	9·2593	9·1129	8·9711	8·8337	8·7004	8·5711	8·4457	8·3238					
7	8·2055	8·0905	7·9787	7·8700	7·7642	7·6613	7·5611	7·4635	7·3684	7·2757					
8	7·1853	7·0972	7·0112	6·9273	6·8454	6·7655	6·6874	6·6111	6·5366	6·4637					
9	6·3925	6·3228	6·2546	6·1880	6·1227	6·0589	5·9963	5·9351	5·8751	5·8164					
10	5·7588	5·7023	5·6470	5·5928	5·5396	5·4874	5·4362	5·3860	5·3367	5·2883					
11	5·2408	5·1942	5·1484	5·1034	5·0593	5·0159	4·9732	4·9313	4·8901	4·8496					
12	4·8097	4·7706	4·7321	4·6942	4·6569	4·6202	4·5841	4·5486	4·5137	4·4793					
13	4·4454	4·4121	4·3792	4·3469	4·3150	4·2837	4·2527	4·2223	4·1923	4·1627					
14	4·1336	4·1048	4·0765	4·0486	4·0211	3·9939	3·9672	3·9408	3·9147	3·8890					
15	3·86370	83871	81404	78970	76568	74198	71858	69549	67269	65018					
16	3·62796	60601	58434	56294	54181	52094	50032	47995	45983	43995					
17	3·42030	40089	38171	36276	34403	32551	30721	28912	27123	25355					
18	3·23607	21878	20169	18479	16808	15155	13520	11903	10303	08721	274	548	822	1047	1371
19	3·07155	05607	04075	02559	01059	2·99574	2·98106	2·96652	2·95213	2·93790	246	492	738	985	1231
20	2·92380	90986	89605	88238	86885	85545	84219	82906	81605	80318	222	445	667	889	1111
21	2·79043	77780	76530	75292	74065	72850	71647	70455	69275	68105	201	403	604	806	1008
22	2·66947	65799	64662	63535	62419	61313	60217	59130	58054	56988	181	367	550	734	918
23	2·55930	54883	53845	52815	51795	50784	49782	48789	47804	46827	168	335	504	671	839
24	2·45859	44900	43948	43005	42070	41142	40222	39311	38406	37509	154	308	462	616	770
25	2·36620	35738	34863	33996	33135	32282	31436	30596	29763	28937	142	283	425	566	708
26	2·28117	27304	26498	25697	24903	24116	23334	22559	21790	21026	130	262	392	523	654
27	2·20269	19517	18771	18031	17297	16568	15845	15127	14414	13707	121	242	363	484	606
28	2·13005	12309	11617	10931	10250	09574	08903	08236	07575	06918	112	225	337	449	561
29	2·06267	05619	04977	04339	03706	03077	02453	01833	01218	00607	104	209	313	417	522
30	2·00000	1·99398	1·98799	1·98205	1·97615	1·97029	1·96448	1·95870	1·95296	1·94726	97	194	292	389	486
31	1·94160	93598	93040	92486	91935	91388	90845	90305	89769	89237	90	182	272	363	454
32	1·88708	88183	87661	87142	86627	86116	85608	85103	84601	84103	85	170	255	340	425
33	1·83608	83116	82627	82142	81659	81180	80704	80231	79761	79293	79	159	239	318	398
34	1·78829	78368	77910	77454	77001	76552	76105	75661	75219	74781	75	149	224	299	374
35	1·74345	73911	73481	73053	72628	72205	71785	71367	70953	70540	70	140	211	281	351
36	1·70130	69723	69318	68915	68515	68117	67722	67329	66938	66550	66	132	198	264	331
37	1·66164	65780	65399	65020	64643	64268	63895	63525	63157	62791	62	124	187	249	311
38	1·62427	62065	61705	61348	60992	60639	60287	59938	59590	59245	59	117	176	235	294
39	1·58902	58560	58221	57883	57547	57213	56881	56551	56223	55897	55	111	166	221	278
40	1·55572	55250	54929	54610	54292	53977	53663	53351	53041	52732	52	105	157	210	262
41	1·52425	52120	51817	51515	51215	50916	50619	50324	50030	49738	49	99	149	198	248
42	1·49448	49159	48871	48586	48301	48019	47738	47458	47180	46903	47	94	141	188	235
43	1·46628	46354	46082	45811	45542	45274	45007	44742	44479	44217	44	89	133	178	223
44	1·43956	43696	43438	43181	42926	42672	42419	42168	41918	41669	42	84	126	169	211

NATURAL COSECANTS

Degrees	0′	6′	12′	18′	24′	30′	36′	42′	48′	54′	Mean Differences				
											1	2	3	4	5
45	1·41421	41175	40930	40687	40444	40203	39963	39725	39487	39251	40	79	120	160	200
46	1·39016	38783	38550	38319	38089	37860	37632	37406	37180	36956	38	76	114	152	190
47	1·36733	36511	36290	36070	35852	35634	35418	35203	34988	34775	36	71	108	145	181
48	1·34563	34352	34142	33934	33726	33519	33314	33109	32905	32703	34	69	103	137	172
49	1·32501	32301	32101	31903	31705	31509	31313	31119	30925	30732	32	65	98	130	163
50	1·30541	30350	30160	29971	29784	29597	29411	29226	29041	28858	31	62	93	124	155
51	1·28676	28495	28314	28134	27956	27778	27601	27425	27250	27075	30	59	89	118	148
52	1·26902	26729	26557	26387	26216	26047	25879	25711	25545	25379	28	56	84	113	141
53	1·25214	25049	24885	24723	24561	24400	24240	24081	23922	23764	27	54	80	107	134
54	1·23607	23450	23295	23140	22986	22833	22680	22528	22377	22227	26	51	77	102	128
55	1·22077	21929	21781	21633	21487	21341	21195	21051	20907	20764	24	49	73	97	121
56	1·20622	20480	20339	20199	20059	19920	19782	19645	19508	19372	23	46	69	92	116
57	1·19236	19102	18967	18834	18701	18569	18437	18307	18176	18047	22	44	66	88	110
58	1·17918	17790	17662	17535	17409	17283	17158	17033	16909	16786	21	42	63	84	105
59	1·16663	16541	16420	16299	16179	16059	15940	15822	15704	15587	20	40	60	80	99
60	1·15470	15354	15239	15124	15009	14896	14782	14670	14558	14446	19	38	57	76	95
61	1·14335	14225	14115	14006	13897	13789	13682	13575	13468	13362	18	36	54	72	90
62	1·13257	13152	13048	12944	12841	12738	12636	12534	12433	12333	17	34	51	68	85
63	1·12233	12133	12034	11936	11838	11740	11643	11547	11451	11355	16	32	49	65	81
64	1·11260	11166	11072	10978	10885	10793	10701	10609	10518	10428	15	31	46	61	77
65	1·10338	10248	10159	10071	09982	09895	09808	09721	09635	09549	15	29	44	58	73
66	1·09464	09379	09294	09211	09127	09044	08962	08880	08798	08717	14	28	41	55	69
67	1·08636	08556	08476	08397	08318	08239	08161	08084	08006	07930	13	26	39	52	64
68	1·07853	07778	07702	07627	07553	07479	07405	07332	07259	07186	12	25	37	49	62
69	1·07115	07043	06972	06901	06831	06761	06691	06622	06554	06486	12	23	35	46	58
70	1·06418	06350	06283	06217	06151	06085	06020	05955	05890	05826	11	22	33	44	55
71	1·05762	05699	05636	05573	05511	05449	05388	05327	05266	05206	10	21	31	41	51
72	1·05146	05087	05028	04969	04911	04853	04795	04738	04682	04625	10	19	29	38	48
73	1·04569	04514	04458	04403	04349	04295	04241	04188	04135	04082	9	18	26	34	44
74	1·04030	03978	03927	03875	03825	03774	03724	03674	03625	03576	8	17	25	33	42
75	1·03528	03479	03432	03384	03337	03290	03244	03197	03152	03106	8	16	23	31	39
76	1·03061	03017	02972	02928	02885	02842	02799	02756	02714	02672	7	14	22	29	36
77	1·02630	02589	02548	02508	02468	02428	02388	02349	02311	02272	7	13	20	26	33
78	1·02234	02196	02159	02122	02085	02049	02013	01977	01941	01906	6	12	18	24	30
79	1·01872	01837	01803	01769	01736	01703	01670	01638	01606	01574	5	11	16	22	27
80	1·01543	01512	01481	01450	01420	01391	01361	01332	01303	01275	5	10	15	20	24
81	1·01247	01219	01191	01164	01137	01111	01084	01059	01033	01008	4	9	13	18	22
82	1·00983	00958	00934	00910	00886	00863	00840	00817	00795	00773	4	8	12	15	19
83	1·00751	00730	00708	00688	00667	00647	00627	00608	00588	00569	3	7	10	13	17
84	1·00551	00533	00515	00497	00480	00463	00446	00429	00413	00397	3	6	8	11	14
85	1·00382	00367	00352	00337	00323	00309	00296	00283	00269	00257	2	5	7	9	12
86	1·00244	00232	00220	00209	00198	00187	00176	00166	00156	00147	2	4	5	7	9
87	1·00137	00128	00120	00111	00103	00095	00088	00081	00074	00067	1	3	4	5	6
88	1·00061	00055	00049	00044	00039	00034	00030	00026	00022	00018	1	2	2	3	4
89	1·00015	00012	00010	00007	00005	00004	00002	00001	00001	00000	0	1	1	1	1

NATURAL SECANTS

Degrees	0′	6′	12′	18′	24′	30′	36′	42′	48′	54′	Mean Differences				
											1	2	3	4	5
0	1·00000	00000	00001	00001	00002	00004	00005	00007	00010	00012	0	1	1	1	1
1	1·00015	00018	00022	00026	00030	00034	00039	00044	00049	00055	1	2	2	3	4
2	1·00061	00067	00074	00081	00088	00095	00103	00111	00120	00128	1	3	4	5	6
3	1·00137	00147	00156	00166	00176	00187	00198	00209	00220	00232	2	4	5	7	9
4	1·00244	00257	00269	00282	00296	00309	00323	00337	00352	00367	2	5	7	9	12
5	1·00382	00397	00413	00429	00446	00463	00480	00497	00515	00533	3	6	8	11	14
6	1·00551	00569	00588	00608	00627	00647	00667	00688	00708	00730	3	7	10	13	17
7	1·00751	00773	00795	00817	00840	00863	00886	00910	00934	00958	4	8	12	15	19
8	1·00983	01008	01033	01059	01084	01111	01137	01164	01191	01219	4	9	13	18	22
9	1·01247	01275	01303	01332	01361	01391	01420	01450	01481	01512	5	10	15	20	24
10	1·01543	01574	01606	01638	01670	01703	01736	01769	01803	01837	5	11	16	22	27
11	1·01872	01906	01941	01977	02013	02049	02085	02122	02159	02196	6	12	18	24	30
12	1·02234	02272	02311	02349	02388	02428	02468	02508	02548	02589	7	13	20	26	33
13	1·02630	02672	02714	02756	02799	02842	02885	02928	02972	03017	7	14	22	29	36
14	1·03061	03106	03152	03197	03244	03290	03337	03384	03432	03479	8	16	23	31	39
15	1·03528	03576	03625	03674	03724	03774	03825	03875	03927	03978	8	17	25	33	42
16	1·04030	04082	04135	04188	04241	04295	04349	04403	04458	04514	9	18	26	34	44
17	1·04569	04625	04682	04738	04795	04853	04911	04969	05028	05087	10	19	29	38	48
18	1·05146	05206	05266	05327	05388	05449	05511	05573	05636	05699	10	21	31	41	51
19	1·05762	05826	05890	05955	06020	06085	06151	06217	06283	06350	11	22	33	44	55
20	1·06418	06486	06554	06622	06691	06761	06831	06901	06972	07043	12	23	35	46	58
21	1·07114	07186	07259	07332	07405	07479	07553	07627	07702	07778	12	25	37	49	62
22	1·07853	07930	08006	08084	08161	08239	08318	08397	08476	08556	13	26	39	52	64
23	1·08636	08717	08798	08880	08962	09044	09127	09211	09294	09379	14	28	41	55	69
24	1·09464	09549	09635	09721	09808	09895	09982	10071	10159	10248	15	29	44	58	73
25	1·10338	10428	10518	10609	10701	10793	10885	10978	11072	11166	15	31	46	61	77
26	1·11260	11355	11451	11547	11643	11740	11838	11936	12034	12133	16	32	49	65	81
27	1·12233	12333	12433	12534	12636	12738	12841	12944	13048	13152	17	34	51	68	85
28	1·13257	13362	13468	13575	13682	13789	13897	14006	14115	14225	18	36	54	72	90
29	1·14335	14446	14558	14670	14782	14896	15009	15124	15239	15354	19	38	57	76	95
30	1·15470	15587	15704	15822	15940	16059	16179	16299	16420	16541	20	40	60	80	99
31	1·16663	16786	16909	17033	17158	17283	17409	17535	17662	17790	21	42	63	84	105
32	1·17918	18047	18176	18307	18437	18569	18701	18834	18967	19102	22	44	66	88	110
33	1·19236	19372	19508	19645	19782	19920	20059	20199	20339	20480	23	46	69	92	116
34	1·20622	20764	20907	21051	21195	21341	21487	21633	21781	21929	24	49	73	97	121
35	1·22077	22227	22377	22528	22680	22833	22986	23140	23295	23450	26	51	77	102	128
36	1·23607	23764	23922	24081	24240	24400	24561	24723	24886	25049	27	54	80	107	134
37	1·25214	25379	25545	25711	25879	26047	26216	26387	26557	26729	28	56	84	113	141
38	1·26902	27075	27250	27425	27601	27778	27956	28134	28314	28495	30	59	89	118	148
39	1·28676	28858	29042	29226	29411	29597	29784	29971	30160	30350	31	62	93	124	155
40	1·30541	30732	30925	31119	31313	31509	31705	31903	32101	32301	32	65	98	130	163
41	1·32501	32703	32905	33109	33314	33519	33726	33934	34142	34352	34	69	103	137	172
42	1·34563	34775	34988	35203	35418	35634	35852	36070	36290	36511	36	71	108	145	181
43	1·36733	36956	37180	37406	37632	37860	38089	38319	38550	38783	38	76	114	152	190
44	1·39016	39251	39487	39725	39963	40203	40444	40687	40930	41175	40	80	120	160	200

NATURAL SECANTS

Degrees	0′	6′	12′	18′	24′	30′	36′	42′	48′	54′	Mean Differences				
											1	2	3	4	5
45	1·41421	41669	41918	42168	42419	42672	42926	43181	43438	43696	42	84	126	169	211
46	1·43956	44217	44479	44742	45007	45274	45542	45811	46082	46354	44	89	133	178	223
47	1·46628	46903	47180	47458	47738	48019	48301	48586	48871	49159	47	94	141	188	235
48	1·49448	49738	50030	50324	50619	50916	51215	51515	51817	52120	49	99	149	198	248
49	1·52425	52732	53041	53351	53663	53977	54292	54610	54929	55250	52	105	157	209	262
50	1·55572	55897	56223	56551	56881	57213	57547	57883	58221	58560	55	111	166	222	278
51	1·58902	59245	59590	59938	60287	60639	60992	61348	61705	62065	59	117	176	235	294
52	1·62427	62791	63157	63525	63895	64268	64643	65020	65399	65780	62	124	187	249	311
53	1·66164	66550	66938	67329	67722	68117	68515	68915	69318	69723	66	132	198	264	331
54	1·70130	70540	70953	71367	71785	72205	72628	73053	73481	73911	70	140	211	281	351
55	1·74345	74781	75219	75661	76105	76552	77001	77454	77910	78368	75	149	224	299	374
56	1·78829	79293	79761	80231	80704	81180	81659	82142	82627	83116	79	159	239	318	398
57	1·83608	84103	84601	85103	85608	86116	86627	87142	87661	88183	85	170	253	340	425
58	1·88708	89237	89769	90305	90845	91388	91935	92486	93040	93598	90	182	272	363	454
59	1·94160	94726	95296	95870	96448	97029	97615	98205	98799	99398	97	194	292	389	486
60	2·0000	00607	01218	01833	02453	03077	03706	04339	04977	05619	104	209	313	417	522
61	2·06267	06918	07575	08236	08903	09574	10250	10931	11617	12309	112	225	339	449	561
62	2·13005	13707	14414	15127	15845	16568	17297	18031	18772	19517	121	242	363	484	606
63	2·20269	21026	21790	22559	23334	24116	24903	25697	26498	27304	130	262	392	523	654
64	2·28117	28937	29763	30596	31436	32282	33135	33996	34863	35738	142	283	425	567	709
65	2·36620	37509	38406	39311	40222	41142	42070	43005	43948	44900	154	308	462	616	770
66	2·45859	46827	47804	48789	49782	50784	51795	52815	53845	54883	168	335	504	672	839
67	2·55930	56988	58054	59130	60217	61313	62419	63535	64662	65799	183	367	550	734	918
68	2·66947	68105	69275	70455	71647	72850	74065	75292	76530	77780	200	403	605	806	1008
69	2·79043	80318	81605	82906	84219	85545	86885	88238	89605	90986					
70	2·92380	93790	95213	96652	98106	99574	3·01059	3·02559	3·04075	3·05607					
71	3·07155	08721	10303	11903	13520	15155	16808	18479	20169	21878					
72	3·23607	25355	27123	28912	30721	32551	34403	36276	38171	40089					
73	3·42030	43995	45983	47995	50032	52094	54181	56294	58434	60601					
74	3·62796	65018	67269	69549	71858	74198	76568	78970	81404	83871					
75	3·86370	88904	91473	94076	96716	99393	4·02107	4·04860	4·07652	4·10484					
76	4·13357	16271	19228	22229	25275	28366	31503	34689	37923	41206					
77	4·44541	47928	51368	54863	58414	62023	65690	69417	73205	77057					
78	4·80973	84956	89007	93128	97320	5·01585	5·05926	5·10344	5·14842	5·19421					
79	5·24084	28833	33671	38600	43622	48740	53958	59277	64701	70234					
80	5·75877	81635	87511	93509	99633	6·05886	6·12273	6·18797	6·25464	6·32279					
81	6·39245	46369	53655	61110	68738	76547	84542	92731	7·01120	7·09717					
82	7·18530	27566	36835	46346	56107	66130	76424	87001	97873	8·09052					
83	8·20551	32384	44566	57113	70041	83367	97111	9·11292	9·25931	9·41052					
84	9·56677	9·7283	9·8955	10·068	10·248	10·433	10·626	10·826	11·034	11·249					
85	11·474	11·707	11·951	12·204	12·469	12·745	13·035	13·337	13·654	13·987					
86	14·336	14·703	15·089	15·496	15·926	16·380	16·862	17·372	17·914	18·492					
87	19·107	19·766	20·471	21·229	22·044	22·926	23·880	24·918	26·050	27·290					
88	28·654	30·161	31·836	33·708	35·815	38·202	40·930	44·077	47·750	52·090					
89	57·299	63·665	71·622	81·853	95·495	114·59	143·24	190·99	286·48	572·96					

NATURAL COTANGENTS

Degrees	0′	6′	12′	18′	24′	30′	36′	42′	48′	54′	Mean Differences				
											1	2	3	4	5
0	∞	572·957	286·478	190·984	143·237	114·589	95·489	81·847	71·615	63·657					
1	57·290	52·081	47·740	44·066	40·917	38·188	35·801	33·694	31·821	30·145					
2	28·636	27·271	26·031	24·898	23·859	22·904	22·022	21·205	20·446	19·740					
3	19·081	18·464	17·886	17·343	16·832	16·350	15·895	15·464	15·056	14·669					
4	14·301	13·951	13·617	13·300	12·996	12·706	12·429	12·163	11·909	11·664					
5	11·4301	11·2048	10·9982	10·7797	10·5789	10·3854	10·1988	10·0187	9·8448	9·6768					
6	9·5144	9·3572	9·2052	9·0579	8·9152	8·7769	8·6427	8·5126	8·3863	8·2636					
7	8·1443	8·0285	7·9158	7·8062	7·6996	7·5958	7·4947	7·3962	7·3002	7·2066					
8	7·1154	7·0264	6·9395	6·8548	6·7720	6·6912	6·6122	6·5350	6·4596	6·3859					
9	6·3138	6·2432	6·1742	6·1066	6·0405	5·9758	5·9124	5·8502	5·7894	5·7297					
10	5·67128	61397	55777	50264	44857	39552	34345	29235	24218	19293					
11	5·14455	09704	05037	00451	4·95945	4·91516	4·87162	4·82882	4·78673	4·74534					
12	4·70463	66458	62518	58641	54826	51071	47374	43735	40152	36623					
13	4·33148	29724	26352	23030	19756	16530	13350	10216	07127	04081					
14	4·01078	3·98117	3·95196	3·92316	3·89474	3·86671	3·83906	3·81177	3·78485	3·75828					
15	3·73205	70616	68061	65538	63048	60588	58160	55761	53393	51053					
16	3·48741	46458	44202	41973	39771	37594	35443	33317	31216	29139					
17	3·27085	25055	23048	21063	19100	17159	15240	13341	11464	09606					
18	3·07768	05950	04152	02372	00611	2·98868	2·97144	2·95437	2·93748	2·92076					
19	2·90421	88783	87161	85555	83965	82391	80833	79289	77761	76247					
20	2·74748	73263	71792	70335	68892	67462	66046	64642	63252	61874					
21	2·60509	59156	57815	56487	55170	53865	52571	51289	50018	48758					
22	2·47509	46270	45043	43825	42618	41421	40235	39058	37891	36733	199	397	596	795	994
23	2·35585	34447	33317	32197	31086	29984	28891	27806	26730	25663	183	366	549	732	916
24	2·24604	23553	22510	21475	20449	19430	18419	17416	16420	15432	169	338	508	677	846
25	2·14451	13477	12511	11552	10600	09654	08716	07785	06860	05942	157	314	471	629	786
26	2·05030	04125	03227	02335	01449	00569	1·99695	1·98828	1·97966	1·97111	146	292	438	584	731
27	1·96261	95417	94579	93746	92920	92098	91282	90472	89667	88867	136	273	409	546	683
28	1·88073	87283	86499	85720	84946	84177	83413	82654	81899	81150	128	255	383	511	639
29	1·80405	79665	78929	78198	77471	76749	76032	75319	74610	73905	120	240	360	481	600
30	1·73205	72509	71817	71129	70446	69766	69091	68419	67752	67088	113	226	339	452	564
31	1·66428	65772	65120	64471	63826	63185	62548	61914	61283	60657	106	213	319	428	533
32	1·60033	59414	58797	58184	57575	56969	56366	55766	55170	54576	101	201	302	403	504
33	1·53986	53400	52816	52235	51658	51084	50512	49944	49378	48816	95	191	286	382	477
34	1·48256	47699	47146	46595	46046	45501	44958	44418	43881	43347	91	181	272	362	453
35	1·42815	42286	41759	41235	40714	40195	39679	39165	38653	38145	86	172	259	345	431
36	1·37638	37134	36633	36134	35637	35142	34650	34160	33673	33187	82	164	247	329	411
37	1·32704	32224	31745	31269	30795	30323	29853	29385	28919	28456	78	157	235	314	392
38	1·27994	27535	27077	26622	26169	25717	25268	24820	24375	23931	75	150	225	300	375
39	1·23490	23050	22612	22176	21742	21310	20879	20451	20024	19599	72	143	216	288	359
40	1·19175	18754	18334	17916	17500	17085	16672	16261	15851	15443	69	138	207	276	343
41	1·15037	14632	14229	13828	13428	13029	12633	12238	11844	11452	66	132	199	265	331
42	1·11061	10672	10285	09899	09514	09131	08749	08369	07990	07613	63	127	191	255	319
43	1·07237	06862	06489	06117	05747	05378	05010	04644	04279	03915	61	123	184	245	307
44	1·03553	03192	02832	02474	02117	01761	01406	01053	00701	00350	58	118	177	237	296

NATURAL COTANGENTS

Degrees	0′	6′	12′	18′	24′	30′	36′	42′	48′	54′	Mean Differences				
											1	2	3	4	5
45	1·00000	99652	99304	98958	98613	98270	97927	97586	97246	96907	57	114	171	229	286
46	0·96569	96232	95897	95562	95229	94896	94565	94235	93906	93578	55	111	165	221	276
47	0·93252	92926	92601	92277	91955	91633	91313	90993	90674	90357	53	107	160	214	268
48	0·90040	89725	89410	89097	88784	88473	88162	87852	87543	87236	52	103	155	207	259
49	0·86929	86623	86318	86014	85710	85408	85107	84806	84507	84208	50	100	151	201	252
50	0·83910	83613	83317	83022	82727	82434	82141	81849	81558	81268	49	98	147	195	244
51	0·80978	80690	80402	80115	79829	79544	79259	78975	78692	78410	47	95	142	190	237
52	0·78129	77848	77568	77289	77010	76733	76456	76180	75904	75629	46	92	139	185	231
53	0·75355	75082	74810	74538	74267	73996	73726	73457	73189	72921	45	90	135	180	225
54	0·72654	72388	72122	71857	71593	71329	71066	70804	70542	70281	44	88	132	176	219
55	0·70021	69761	69502	69243	68985	68728	68471	68215	67960	67705	43	86	129	171	214
56	0·67451	67197	66944	66692	66440	66189	65938	65688	65438	65189	42	84	126	167	209
57	0·64941	64693	64446	64199	63953	63707	63462	63217	62973	62730	41	82	123	164	205
58	0·62487	62245	62003	61761	61520	61280	61040	60801	60562	60324	40	79	120	160	200
59	0·60086	59849	59612	59376	59140	58905	58670	58435	58201	57968	39	78	118	157	196
60	0·57735	57503	57271	57039	56808	56577	56347	56117	55888	55659	38	77	115	154	192
61	0·55431	55203	54975	54748	54522	54296	54070	53844	53620	53395	38	75	113	151	188
62	0·53171	52947	52724	52501	52279	52057	51835	51614	51393	51173	37	74	111	148	185
63	0·50953	50733	50514	50295	50076	49858	49640	49423	49206	48989	36	73	109	145	182
64	0·48773	48557	48342	48127	47912	47698	47483	47270	47056	46843	36	71	107	143	179
65	0.46631	46418	46206	45995	45784	45573	45362	45152	44942	44732	35	70	105	141	176
66	0·44523	44314	44105	43897	43689	43481	43274	43067	42860	42654	35	69	104	138	173
67	0·42447	42242	42036	41831	41626	41421	41217	41013	40809	40606	34	68	102	136	170
68	0·40403	40200	39997	39795	39593	39391	39190	38988	38787	38587	34	67	101	134	168
69	0·38386	38186	37986	37787	37588	37388	37190	36991	36793	36595	33	66	99	133	166
70	0·36397	36199	36002	35805	35608	35412	35216	35019	34824	34628	33	66	98	131	164
71	0·34433	34238	34043	33848	33654	33460	33266	33072	32878	32685	32	65	97	129	162
72	0·32492	32299	32106	31914	31722	31530	31338	31147	30955	30764	32	64	96	128	160
73	0·30573	30382	30192	30001	29811	29621	29432	29242	29053	28864	32	63	95	127	158
74	0·28675	28486	28297	28109	27920	27732	27545	27357	27169	26982	31	63	94	125	157
75	0·26795	26608	26421	26235	26048	25862	25676	25490	25304	25118	31	62	93	124	155
76	0·24933	24747	24562	24377	24193	24008	23823	23639	23455	23271	31	61	93	124	155
77	0·23087	22903	22719	22536	22353	22169	21986	21804	21621	21438	31	61	92	122	153
78	0·21256	21073	20891	20709	20527	20345	20164	19982	19801	19619	30	60	91	121	152
79	0·19438	19257	19076	18895	18714	18534	18353	18173	17993	17813	30	60	90	120	150
80	0·17633	17453	17273	17093	16914	16734	16555	16376	16196	16017	30	60	90	120	150
81	0·15838	15660	15481	15302	15124	14945	14767	14588	14410	14232	30	59	89	119	149
82	0·14054	13876	13698	13521	13343	13165	12988	12810	12633	12456	30	59	89	118	148
83	0·12278	12101	11924	11747	11570	11394	11217	11040	10863	10687	29	59	88	118	147
84	0·10510	10334	10158	09981	09805	09629	09453	09277	09101	08925	29	59	88	117	147
85	0·08749	08573	08397	08221	08046	07870	07695	07519	07344	07168	29	59	88	117	146
86	0·06993	06817	06642	06467	06291	06116	05941	05766	05591	05416	29	58	88	117	146
87	0·05241	05066	04891	04716	04541	04366	04191	04016	03842	03667	29	58	87	117	146
88	0·03492	03317	03143	02968	02793	02619	02444	02269	02095	01920	29	58	87	116	146
89	0·01746	01571	01396	01222	01047	00873	00698	00524	00349	00175	29	58	87	116	146

CHARACTERISTICS OF INSULATING MATERIALS (25°C)

Material	Dielectric constant (*K*)	Power Factor					Dielectric strength Volts/.001 in.	Specific resistance (Ω/cm. cube)
		60 c/s	1 Kc/s	1 Mc/s	10 Mc/s	100 Mc/s		
Air (normal pressure)	1.0						19.8-22.8	
Aniline formaldehyde resin	3-5	1-6					400	10^{12}
Bakelite (Phenolaldehyde[a] resins)								
no filler (pure)	5			1			400-475	1.5×10^{12}
black moulded	5-5.5			3.5			400-500	
cotton fabric base	4.5-7			2-8			150-500	10^{10}-10^{12}
glass fabric base	3.7-4.5			1-2				
paper base (laminated)	3.8-5.5			2-8			650-750	10^{10}-10^{13}
mica filled	5-6			0.8-1			475-600	
Beeswax	2.9-3.2							
Casein plastics[b]	6.1-6.4			5.2-6			165	
Castor oil	4.3-4.7			7			380	
Celluloid	4-16			5-10				
Cellulose acetate[c]	6-8	3-6	4-6	4-6	5.5		300-1000	4.5×10^{10}
Cellulose nitrate[d]	4-7			2.8-5			300-780	2-30×10^{10}
Ceresin wax	2.5-2.6			0.12-0.21				
Empire cloth (varnished cloth)	2-2.5			2-3			440-550	
Ethyl cellulose	2-2.7	0.7	1.2	1.5			1500	10^{15}
Fibre (red)	2.5-5						150-180	5×10^{9}
Glass								
crown	6.2-7		1	1			500	
flint	7-10		0.45	0.4				
plate	6.8-7.6			0.6-0.8				2×10^{13}
pyrex	4.2-4.9		0.5	0.7		0.54	335	10^{14}
window	7.6-8			1.4			200-250	
Gutta Percha	2.5-4.9		1-3				200-500	5×10^{14} or 10^{15}
Methyl methacrylate resins[e]	2.5-3	7	5	1.5-3	1.9		500	10^{15}
Mica								
clear India	6.4-7.5	2	2	2	2		600-1500	
stove	2.5-8	0.2	0.3	0.2-6	0.02		1250-5500	2×10^{17}
Micalex	7.4			0.18			250	10^{13}
Nylon	3.6			2.2				10^{13}
Paper	2.0-2.6						1250	

CHARACTERISTICS OF INSULATING MATERIALS (25°C) (continued)

Material	Dielectric constant (K)	Power Factor					Dielectric strength Volts/.001 in.	Specific resistance (Ω/cm. cube)
		60 c/s	1 Kc/s	1 Mc/s	10 Mc/s	100 Mc/s		
Paraffin wax (solid)	1.9-2.6			0.1-0.3			300	10^{15}-10^{19}
Polyethylene	2.2			0.06			600	10^{17}
Polystyrene[f]	2.4-2.9 (2.6)	0.02	0.018	0.02	0.02	0.02	500-2500	10^{20}
Porcelain (dry process)	6.2-7.5			0.7-15			40-100	5×10^{8}
Porcelain (wet process)	6.5-7			0.6			150	
Pressboard (untreated)	2.9-4.5						125-300	
Pressboard (oiled)	5						750	
Quartz (fused)	3.5-(3.8)	0.01	0.01	0.015-0.03	0.01	0.05	200	10^{14}-10^{18}
Rubber (hard) or Ebonite	2-3.5 (3)			0.5-1			450	10^{12}-10^{15}
Shellac	2.5-4			0.09			900	10^{16}
Steatite (commercial grade)[g]	4.9-6.5	0.02	0.2	0.2	0.4	0.5		
Steatite (low-loss grade)	4.4	0.02	0.2	0.2	0.18	0.13	150-315	10^{14}-10^{15}
Titanium dioxide[h]	90-170		0.1	0.1				
Urea formaldehyde[i]	5-7	3-5	2-3	2-4	4		300-550	10^{12}-10^{13}
Wire coverings:								
Enamel, ordinary							500-750	
Enamel plastic							4500	
Cotton							140	
Silk							450	
Wood (dry oak)	2.5-6.8 (3)		3.8	4.2				
Wood (paraffined maple)	4.1						115	

a. Phenolaldehyde resins include Acrolite, Bakelite, Catalin, Celeron, Dielecto, Durez, Durite, Formica, Gemstone, Heresite, Indur, Makalot, Marblette, Micarte, Opalon, Prystal, Resinox, Synthane, Textolite, etc.

b. Casein plastics include Aladdinite, Ameroid, Galalith, Erinoid, Lactoid, etc.

c. Cellulose acetate includes Fibestos, Lumerith, Nixonite, Plastacele, Tenite, etc.

d. Cellulose nitrate includes Amerith, Nitron, Nixonoid, Pyralin, etc.

e. Methyl methacrylate resins include Acryloid, Lucite, Perspex, Plexiglas, etc.

f. Polystrene includes Amphenol 912*A*, Distrene, Intelin *IN*45, Loalin, Lustron, Quartz *Q*, Rexoglas, Rhodolene *M*, Ronilla *L*, Styraflex, Styron, Trolitul, Victron, etc.

g. Steatite (Soapstone) includes Alberene, Alsimag, Isolantite, Lava, etc.

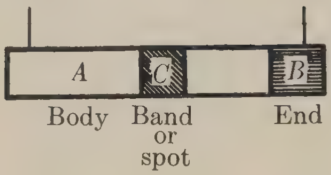
h. Titanium dioxide (Rutile) is used in low temperature-coefficient fixed capacitors

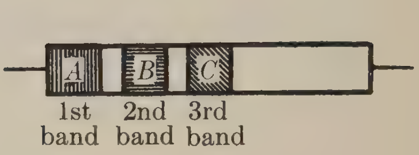
i. Urea formaldehyde resins include Aldur, Beetle, Plaskon, Pollopas, Prystal, etc.

APPENDIX X

COLOR CODES

1. Resistors.

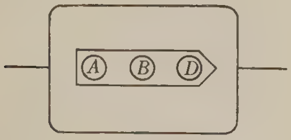




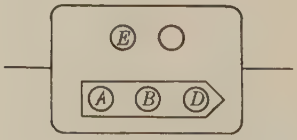
Resistance in Ohms

COLOR	SIGNIFICANT FIGURE		DECIMAL MULTIPLIER	TOLERANCE
	A	*B*	*C*	*D*
Black	0	0	1	
Brown	1	1	10	
Red	2	2	100	
Orange	3	3	1,000	
Yellow	4	4	10,000	
Green	5	5	100,000	
Blue	6	6	1,000,000	
Violet	7	7	10,000,000	
Grey	8	8	100,000,000	
White	9	9	1,000,000,000	
Gold			0.1	5%
Silver			0.01	10%
None				20%

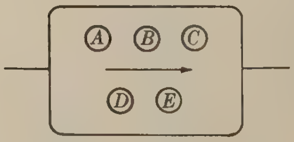
2. Mica capacitors.



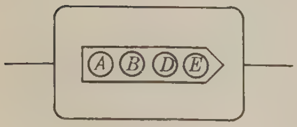
(*a*)



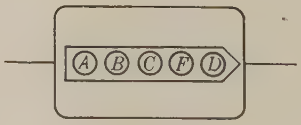
(*b*)



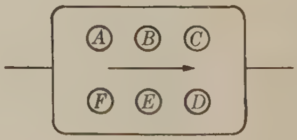
(*c*)



(*b*)



(*d*)

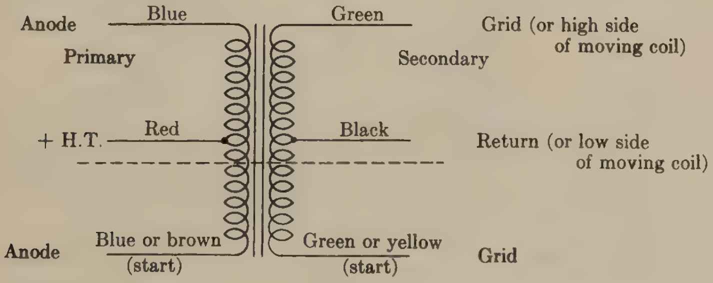


(*f*)

Capacitance in Micromicrofarads

DOT COLOR	SIGNIFICANT FIGURES A	B	C	DECIMAL MULTIPLIER D	TOLERANCE E	VOLTAGE RATING F	DOT COLOR
Black	0	0	0	1			Black
Brown	1	1	1	10	1%	100	Brown
Red	2	2	2	100	2%	200	Red
Orange	3	3	3	1,000	3%	300	Orange
Yellow	4	4	4	10,000	4%	400	Yellow
Green	5	5	5	100,000	5%	500	Green
Blue	6	6	6	1,000,000	6%	600	Blue
Violet	7	7	7	10,000,000	7%	700	Violet
Grey	8	8	8	100,000,000	8%	800	Grey
White	9	9	9	1,000,000,000	9%	900	White
Gold				0.1	5%	1,000	Gold
Silver				0.01	10%	2,000	Silver
None					20%	500	None

3. A-f transformers.



The portion above dotted line applies to single primary and/or secondary transformers

Blue —anode (finish) lead of primary

Red —+H.T. lead

Brown—anode (start) lead on centre tapped primaries (blue may be used for this lead if polarity is not important)

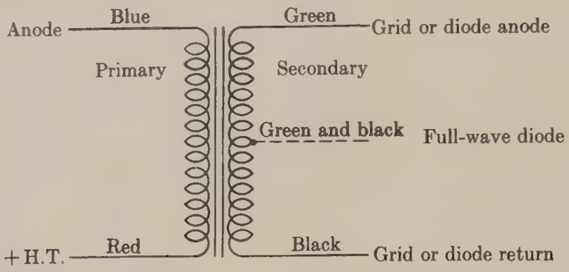
Green —grid (finish) lead to secondary

Black —grid return (this applies whether the secondary is plain or centre tapped)

Yellow—grid (start) lead on centre tapped secondaries (green may be used for this lead if polarity is not important)

Note: These markings apply also to line-to-grid, and valve-to-line transformers.

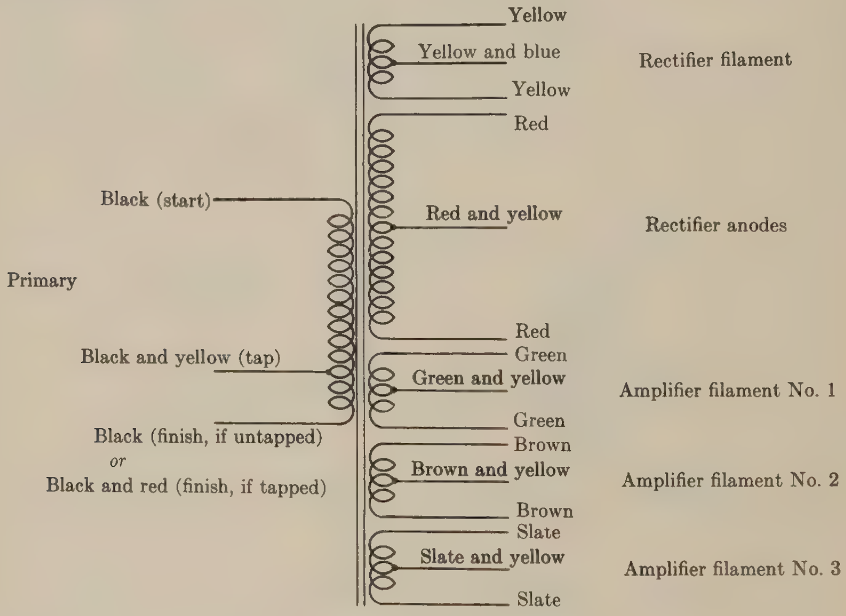
4. I-f transformers.



Blue —anode lead
Red —+H.T. lead
Green —grid (or diode) lead
Black —grid (or diode) return

Note: If the secondary of the i-f transformer is centre tapped, the second diode anode lead is green and black striped, and black is used for the centre tap lead.

5. Power transformers.



Black —primary leads; common, if tapped
Black and yellow—tap
Black and red —finish, if tapped
Red —high voltage winding; rectifier anode winding
Red and yellow —centre tap
Yellow —rectifier filament winding
Yellow and blue —centre tap
Green —amplifier filament No. 1
Green and yellow—centre tap
Brown —amplifier filament No. 2
Brown and yellow—centre tap
Slate —amplifier filament No. 3
Slate and yellow —centre tap

6. Loudspeaker.

Voice coils: Green—finish
Black—start
Field coils: Black and red—start
Yellow and red—finish
Slate and red—tap (if any)

7. Colour system used in Service circuit diagrams

Black —Cathodes which are not at H.T.
Switches, Keys, Jacks, Headphones, Speakers, Microphones and Relays.
Cores of transformers.
Arrows of variable capacitors, variometers, couplings.
Outline of boards, panels, etc. not acting as screens.
Cathodes of indirectly heated valves.
Blue —Aerial circuits.
Negative busbars and negative of main d-c supplies.
Heater circuits of indirectly heated valves.
Red —H.T. supplies.
Anode circuits and screen grid circuits.
Cathode of rectifier valve (at H.T.).
Positive busbars and positive of main d-c supplies.
Green —Grid circuits.
Brown—A-f output circuit.
A-c low tension circuits.
Secondary of induction coil.
Microphone circuits.
Orange—D-c and auxiliary circuits from busbars.
Violet —R-f circuits and by-pass capacitors.
Closed oscillatory circuits.
Crystals.
Yellow—Screens.

STANDARD RESISTOR VALUES

	Ohms						Megohms			Tolerance		
RMA No.	0 to 1.0	1 to 9.1	10 to 91	100 to 910	1,000 to 9,100	10,000 to 91,000	0.1 to 0.91	1.0 to 9.1	10 to 91	5%	10%	20%
10		1.0	10	100	1,000	10,000	0.1	1.0	10	✓	✓	✓
11		1.1	11	110	1,100	11,000	0.11	1.1	11	✓		
12		1.2	12	120	1,200	12,000	0.12	1.2	12	✓	✓	
13		1.3	13	130	1,300	13,000	0.13	1.3	13	✓		
15		1.5	15	150	1,500	15,000	0.15	1.5	15	✓	✓	✓
16		1.6	16	160	1,600	16,000	0.16	1.6	16	✓		
18		1.8	18	180	1,800	18,000	0.18	1.8	18	✓	✓	
20		2.0	20	200	2,000	20,000	0.20	2.0	20	✓		
22		2.2	22	220	2,200	22,000	0.22	2.2	22	✓	✓	✓
24	0.24	2.4	24	240	2,400	24,000	0.24	2.4	24	✓		
	0.25		25	250	2,500	25,000	0.25	2.5	25			
27	0.27	2.7	27	270	2,700	27,000	0.27	2.7	27	✓	✓	
30	0.30	3.0	30	300	3,000	30,000	0.30	3.0	30	✓		
33	0.33	3.3	33	330	3,300	33,000	0.33	3.3	33	✓	✓	✓
36	0.36	3.6	36	360	3,600	36,000	0.36	3.6	36	✓		
39	0.39	3.9	39	390	3,900	39,000	0.39	3.9	39	✓	✓	
43	0.43	4.3	43	430	4,300	43,000	0.43	4.3	43	✓		
47	0.47	4.7	47	470	4,700	47,000	0.47	4.7	47	✓	✓	✓
	0.50	5.0	50	500	5,000	50,000	0.50	5.0	50			
51	0.51	5.1	51	510	5,100	51,000	0.51	5.1	51	✓		
56	0.56	5.6	56	560	5,600	56,000	0.56	5.6	56	✓	✓	
62	0.62	6.2	62	620	6,200	62,000	0.62	6.2	62	✓	✓	✓
68	0.68	6.8	68	680	6,800	68,000	0.68	6.8	68	✓		
75	0.75	7.5	75	750	7,500	75,000	0.75	7.5	75	✓	✓	
82	0.82	8.2	82	820	8,200	82,000	0.82	8.2	82	✓		
91	0.91	9.1	91	910	9,100	91,000	0.91	9.1	91	✓		

SPECIFIC RESISTANCES

Material (at 20°C)	Ohms per circular mil-foot
"Advance" (alloy)	294.
Aluminium	17.
Bismuth	663.
Brass	42.
Carbon (coke, lampblack)	22,000.
"Climax" (alloy)	480.
"Constantan" (alloy)	294.
Copper (annealed)	10.35
Copper (hard drawn)	10.60
Excello (alloy)	552.
German Silver (18%)	198.
German Silver (30%) (Constantan)	294.
Gold	14.6
Graphite	4,300.
Iron (pure, annealed)	61.
Iron (cast)	435.
Ia Ia, hard (alloy)	300.
Lead	132.35
Manganin	264.
Mercury	576.
Molybdenum (drawn)	34.
Monel metal (alloy)	252.
Nichrome (alloy)	675.
Nickel	60.
Phosphor-bronze	23.7
Platinum	60.
Silver	9.56
Steel (cast)	115.
Steel (soft, carbon)	96.
Steel (transformer)	66.
Tantalum	93.
Therlo (alloy)	282.
Tin	69.
Tungsten (drawn)	34.
Zinc	35.

COPPER WIRE TABLE

Gauge B. & S.	No. Nearest British S.W.G.	Diameter		Area	Resistance Ohms per 1000 ft. 25°C	Weight Pounds per 1000 ft.	Allowable Current ② Amperes			Turns per Linear Inch ①				Gauge Nearest British S.W.G.	No. B. & S.
		mils	mm.	Circular mils			Rubber Insul.	Varn. Cambric Insul.	Other Insulations	Enamel	S.C.C.	D.S.C. or S.C.C.	D.C.C.		
0000		460.0		2.116×10^5	0.0500	641.0	225	270	325						0000
000		410.0		1.678×10^5	0.0630	508.0	175	210	275						000
00		365.0		1.331×10^5	0.0795	403.0	150	180	225						00
0		325.0		1.055×10^5	0.1000	319.0	125	150	200						0
1	1	289.3	7.348	8.369×10^4	0.1264	253.0	100	120	150			3.3	3.3	1	1
2	3	259.6	6.544	6.637×10^4	0.1593	201.0	90	110	125			3.8	3.6	3	2
3	4	229.4	5.827	5.264×10^4	0.2009	159.0	80	95	100			4.2	4.0	4	3
4	5	204.3	5.189	4.174×10^4	0.2533	126.0	70	85	90			4.7	4.5	5	4
5	7	181.9	4.621	3.310×10^4	0.3195	100.0	55	65	80			5.2	5.0	7	5
6	8	162.0	4.115	2.625×10^4	0.4028	79.5	50	60	70			5.9	5.6	8	6
7	9	144.3	3.665	2.082×10^4	0.5080	63.0						6.5	6.2	9	7
8	10	128.5	3.264	1.651×10^4	0.6405	50.0	35	40	50	7.6		7.4	7.1	10	8
9	11	114.4	2.906	1.309×10^4	0.8077	39.6				8.6		8.2	7.8	11	9
10	12	101.9	2.588	1.038×10^4	1.018	31.4	25	30	30	9.6		9.3	8.9	12	10
11	13	90.74	2.305	8.234×10^3	1.284	24.9				10.7		10.3	9.8	13	11
12	14	80.81	2.053	6.530×10^3	1.619	19.8	20	25	25	12.0		11.5	10.9	14	12
13	15	71.96	1.828	5.178×10^3	2.042	15.7				13.5		12.8	12.0	15	13
14	16	64.08	1.628	4.107×10^3	2.575	12.4	15	18	20	15.0		14.2	13.8	16	14
15	17	57.07	1.450	3.257×10^3	3.247	9.86				16.8		15.8	14.7	17	15
16	18	50.82	1.291	2.583×10^3	4.094	7.82	6			18.9	18.9	17.9	16.4	18	16
17	18	45.26	1.150	2.048×10^3	5.163	6.20				21.2	21.2	19.9	18.1	18	17
18	19	40.30	1.024	1.624×10^3	6.510	4.92	3			23.6	23.6	22.0	19.8	19	18
19	20	35.89	0.9116	1.288×10^3	8.210	3.90				26.4	26.4	24.4	21.8	20	19

COPPER WIRE TABLE (Continued)

Gauge B. & S.	No. Nearest British S.W.G.	Diameter		Area	Resistance Ohms per 1000 ft. 25°C	Weight Pounds per 1000 ft.	Allowable Currents Amperes			Turns per Linear Inch —				Gauge Nearest British S.W.G.	No. B. & S.
		mils	mm.	Circular mils			Rubber Insul.	Varn. Cambric Insul.	Other Insul-ations	Enamel	S.C.C.	D.S.C. or S.C.C.	D.C.C.		
20	21	31.96	0.8118	1.022×10³	10.35	3.09				29.4	29.4	27.0	23.8	21	20
21	22	28.46	0.7230	810.1	13.05	2.45				33.1	32.7	29.8	26.0	22	21
22	23	25.35	0.6438	642.4	16.46	1.95				37.0	36.5	34.1	30.0	23	22
23	24	22.57	0.5733	509.5	20.76	1.54				41.3	40.6	37.6	31.6	24	23
24	25	20.10	0.5106	404.0	26.17	1.22				46.3	45.3	41.5	35.6	25	24
25	26	17.90	0.4547	320.4	33.00	0.970				51.7	50.4	45.6	38.6	26	25
26	27	15.94	0.4049	254.1	41.62	0.769				58.0	55.6	50.2	41.8	27	26
27	29	14.20	0.3606	201.5	52.48	0.610				64.9	61.5	55.0	45.0	29	27
28	30	12.64	0.3211	159.8	66.17	0.484				72.7	68.6	60.2	48.5	30	28
29	31	11.26	0.2859	126.7	83.44	0.384				81.6	74.8	65.4	51.8	31	29
30	33	10.03	0.2546	100.5	105.2	0.304				90.5	83.3	71.5	55.5	33	30
31	34	8.928	0.2268	79.70	132.7	0.241				101.	92.0	77.5	59.2	34	31
32	36	7.950	0.2019	63.21	167.3	0.191				113.	101.	83.6	62.6	36	32
33	37	7.080	0.1798	50.13	211.0	0.152				127.	110.	90.3	66.3	37	33
34	38	6.305	0.1601	39.75	266.0	0.120				143.	120.	97.0	70.0	38	34
35	38–39	5.615	0.1426	31.52	335.0	0.0954				158.	132.	104.	73.5	38–39	35
36	39–40	5.000	0.1270	25.00	423.0	0.0757				175.	143.	111.	77.0	39–40	36
37	41	4.453	0.1131	19.83	533.4	0.0600				194.	154.	118.	80.3	41	37
38	42	3.965	0.1007	15.72	672.6	0.0476				224.	166.	126.	83.6	42	38
39	43	3.531	0.897	12.47	848.1	0.0377				248.	181.	133.	86.6	43	39
40	44	3.145	0.0799	9.88	1069.	0.0299				282.	194.	140.	89.7	44	40

① Figures given are approximate only, since the thickness of the insulation used varies with different manufacturers.
② National Electrical Code.

APPENDIX XIV

UNITS OF LENGTH

1 kilometre (km.) = 1,000 metres = .6214 miles
1 decimetre (dm.) = 0.1 metre
1 centimetre (cm.) = 0.01 metre
1 millimetre (mm.) = 0.001 metre
1 micron (μ) = 0.000,001 metre
= 0.001 millimetre
1 statute mile = 1,760 yards = 5,280 feet
= 1,609.3 metres

1 foot = $\frac{1}{3}$ yard = 12 inches
1 inch = $\frac{1}{36}$ yard = $\frac{1}{12}$ foot = 2.540 cm.
1 mil = $\frac{1}{1000}$ inch

$$\left.\begin{array}{l}\text{1 nautical mile}\\ \text{1 Admiralty mile}\end{array}\right\} = \text{1,853.2 metres} = \text{6080.0 feet} = \text{1.1515 statute miles}$$

$$\left.\begin{array}{l}\text{1 U.S. nautical mile}\\ \text{1 sea mile}\\ \text{1 geographical mile}\end{array}\right\} = \text{1,853.248 metres} = \text{6080.20 feet}$$

1 international nautical mile = 1,852 metres = 6076.10 feet
1 league = 3 nautical miles
1 fathom = 6 feet

APPENDIX XV

FREQUENCY – WAVELENGTH
LC VALUES

FREQ. (f) (Kc/s)	WAVE-LENGTH (λ) (m)	LC ($\mu H-\mu F$)
300	1000	2.816×10^{-1}
350	857	2.068×10^{-1}
400	750	1.583×10^{-1}
450	666	2.251×10^{-1}
500	600	1.014×10^{-1}
550	545	8.375×10^{-2}
600	500	7.037×10^{-2}
650	461	5.997×10^{-2}
700	429	5.170×10^{-2}
750	400	4.502×10^{-2}
800	375	3.958×10^{-2}
850	353	3.507×10^{-2}
900	333	3.127×10^{-2}
950	316	2.807×10^{-2}
1,000	300	2.533×10^{-2}
1,050	286	2.298×10^{-2}
1,100	273	2.094×10^{-2}
1,150	261	1.916×10^{-2}
1,200	250	1.759×10^{-2}
1,250	240	1.622×10^{-2}
1,300	231	1.499×10^{-2}
1,350	222	1.390×10^{-2}
1,400	214	1.292×10^{-2}
1,450	207	1.205×10^{-2}
1,500	200	1.126×10^{-2}
1,600	187	9.896×10^{-3}
1,700	176	8.766×10^{-3}
1,800	167	7.820×10^{-3}
1,900	158	7.019×10^{-3}

FREQ. (f) (Mc/s)	WAVE-LENGTH (λ) (m)	LC ($\mu H-\mu F$)
2	150	6.333×10^{-3}
3	100	2.816×10^{-3}
4	75.0	1.583×10^{-3}
5	60.0	1.014×10^{-3}
6	50.0	7.037×10^{-4}
7	42.9	5.170×10^{-4}
8	37.5	3.958×10^{-4}
9	33.3	3.127×10^{-4}
10	30.0	2.533×10^{-4}
11	27.3	2.094×10^{-4}
12	25.0	1.759×10^{-4}
13	23.1	1.499×10^{-4}
14	21.4	1.292×10^{-4}
15	20.0	1.126×10^{-4}
16	18.7	9.896×10^{-5}
17	17.6	8.766×10^{-5}
18	16.7	7.820×10^{-5}
19	15.8	7.019×10^{-5}
20	15.0	6.333×10^{-5}
21	14.3	5.745×10^{-5}
22	13.6	5.235×10^{-5}
23	13.0	4.789×10^{-5}
24	12.5	4.398×10^{-5}
25	12.0	4.053×10^{-5}
26	11.5	3.748×10^{-5}
27	11.1	3.475×10^{-5}
28	10.7	3.231×10^{-5}
29	10.3	3.012×10^{-5}
30	10.0	2.816×10^{-5}

Conversions for frequencies between 30 and 30,000 Mc/s may be found from this table by using appropriate multipliers.

For 30-300 Mc/s read corresponding values of 300 Kc/s—3 Mc/s from the table and convert by wavelength $\times10^{-2}$; $LC\times10^{-4}$.

For 300-30,000 Mc/s read corresponding values of 300 Kc/s—30 Mc/s from the table and convert by wavelength $\times10^{-3}$; $LC\times10^{-5}$.

APPENDIX XVI

REACTANCE-FREQUENCY CHARTS

In the nomographs on pages A:31-A:33, inductance, capacitance, reactance and frequency are plotted on a series of vertical lines, one for each variable. The frequency range plotted comprises the frequency spectrum from 1 c/s to 1000 Mc/s. To make it possible to plot actual magnitudes on the other scales the frequency spectrum is divided into the following three parts:

Chart I (page A:31)—covers the frequency range from 1 c/s to 1000 c/s
Chart II (page A:32)—covers the frequency range from 1 Kc/s to 1000 Kc/s
Chart III (page A:33)—covers the frequency range from 1 Mc/s to 1000 Mc/s

1. To Find Reactance.

The reactance offered by an inductance or capacitance at any frequency may be determined by placing a straight-edge across the chart to connect the known quantities.

Example. Find the capacitive reactance of a $0.01\mu F$ capacitor to a 400 c/s current.

Place a straight-edge across Chart I to connect the points $0.01\mu F$ and 400 c/s.

Read the capacitive reactance ($40{,}000\,\Omega$) from the point at which the straight-edge intersects the reactance scale.

2. To Find the Resonant Frequency.

The charts may be used to find the resonant frequency of a circuit which has any combination of L and C.

Example. Find the resonant frequency of a circuit in which the inductance is $250\mu H$ and the capacitance is $300\mu\mu F$.

Place a straight-edge across Chart II to connect the points $250\mu H$ and $300\mu\mu F$.

Read the resonant frequency (600 Kc/s) from the point at which the straight-edge intersects the reactance scale.

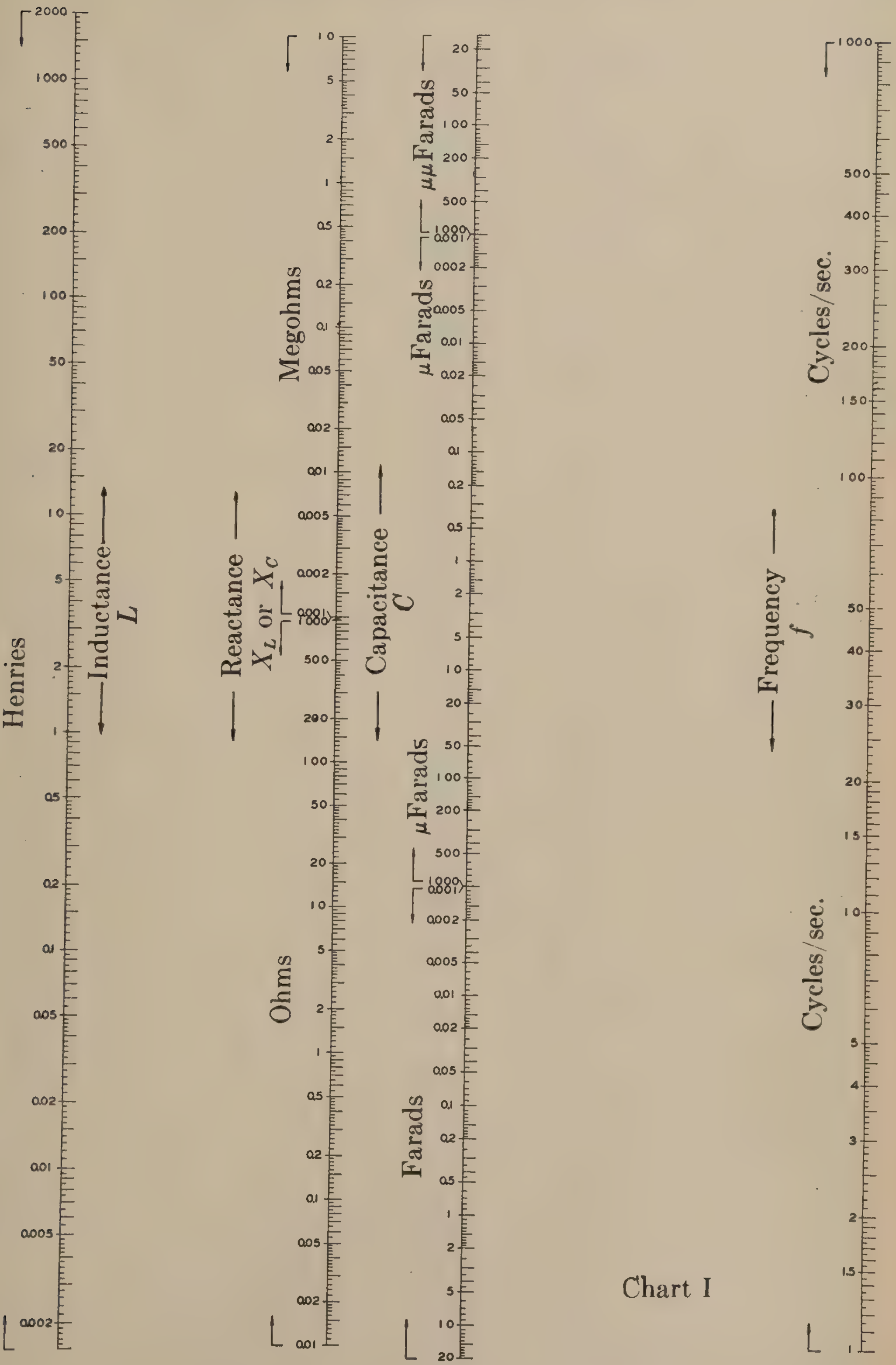

Henries
Inductance
L
2000
1000
500
200
100
50
20
10
5
2
1
0.5
0.2
0.1
0.05
0.02
0.01
0.005
0.002
Megohms
Ohms
Reactance
X_L or X_C
10
5
2
1
0.5
0.2
0.1
0.05
0.02
0.01
0.005
0.002
0.001
1000
500
200
100
50
20
10
5
2
1
0.5
0.2
0.1
0.05
0.02
0.01
Capacitance
C
μμFarads
μFarads
Farads
20
50
100
200
500
1000
0.001
0.002
0.005
0.01
0.02
0.05
0.1
0.2
0.5
1
2
5
10
20
50
100
200
500
1000
0.001
0.002
0.005
0.01
0.02
0.05
0.1
0.2
0.5
1
2
5
10
20
Frequency
f
Cycles/sec.
Cycles/sec.
1000
500
400
300
200
150
100
50
40
30
20
15
10
5
4
3
2
1.5
1

Chart I

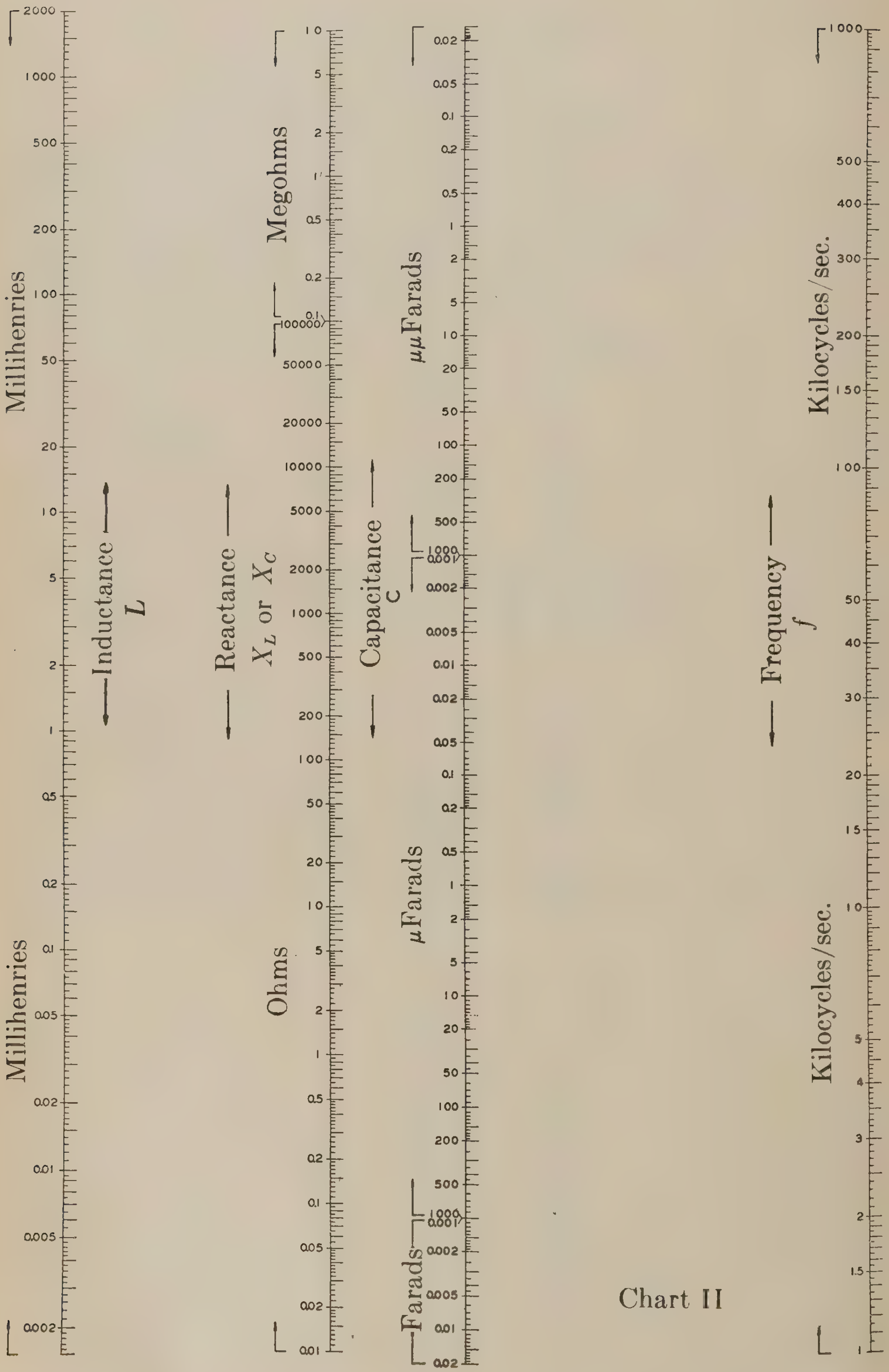

Millihenries
2000
1000
500
200
100
50
20
10
5
2
1
0.5
0.2
0.1
0.05
0.02
0.01
0.005
0.002
Millihenries
Inductance
L
Megohms
10
5
2
1
0.5
0.2
0.1
100000
50000
20000
10000
5000
2000
1000
500
200
100
50
20
10
5
2
1
0.5
0.2
0.1
0.05
0.02
0.01
Ohms
Reactance
X_L or X_C
μμFarads
0.02
0.05
0.1
0.2
0.5
1
2
5
10
20
50
100
200
500
1000
0.001
0.002
0.005
0.01
0.02
0.05
0.1
0.2
0.5
1
2
5
10
20
50
100
200
500
1000
0.001
0.002
0.005
0.01
0.02
μFarads
Farads
Capacitance
C
Kilocycles/sec.
1000
500
400
300
200
150
100
50
40
30
20
15
10
5
4
3
2
1.5
1
Kilocycles/sec.
Frequency
f

Chart II

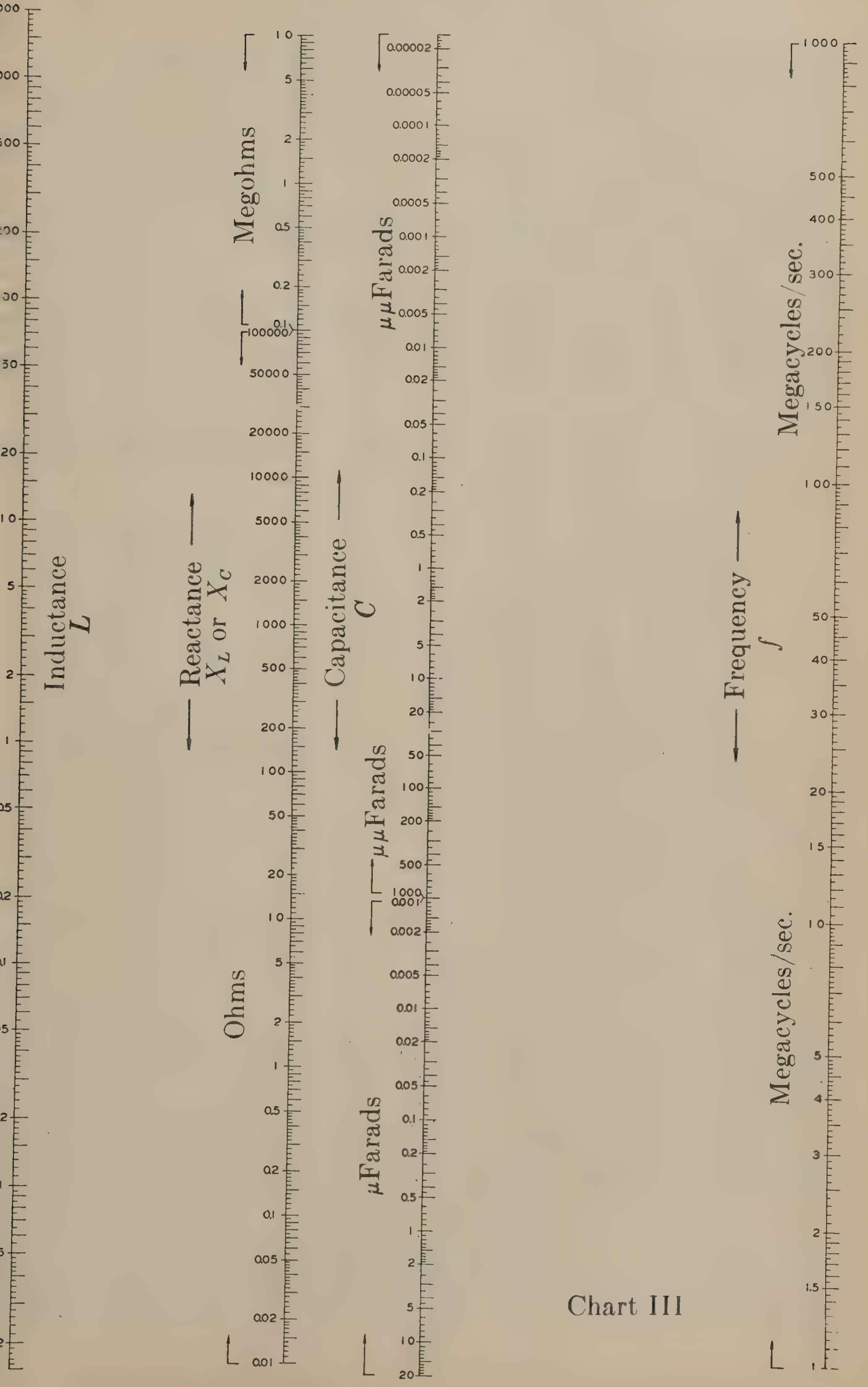

Inductance
L
Reactance
X_L or X_C
Megohms
10
5
2
1
0.5
0.2
0.1
100000
50000
20000
10000
5000
2000
1000
500
200
100
50
20
10
5
2
1
0.5
0.2
0.1
0.05
0.02
0.01
Ohms
Capacitance
C
μμFarads
0.00002
0.00005
0.0001
0.0002
0.0005
0.001
0.002
0.005
0.01
0.02
0.05
0.1
0.2
0.5
1
2
5
10
20
50
100
200
500
1000
μFarads
0.001
0.002
0.005
0.01
0.02
0.05
0.1
0.2
0.5
1
2
5
10
20
Frequency
f
Megacycles/sec.
1000
500
400
300
200
150
100
50
40
30
20
15
10
5
4
3
2
1.5
1
20
10
5
2
1

Chart III

APPENDIX XVII

DECIBEL CONVERSION TABLE

1. **To convert voltage or current ratios to db.** Read directly from the table.
2. **To convert db to voltage or current ratios.**
   - (a) Locate the db value in the body of the table.
   - (b) Read the required ratio from the left hand column and top row.
3. **To convert power ratios to db.** Since a power ratio equals the square root of the voltage or current ratio the logarithm of the power ratio equals one-half of the logarithm of the voltage or current ratio. Hence, use the table as directed for 1 above and take one-half of the result.

   Example: Given a power ratio of 3.41, find the db value.

   Method: From the table 3.41 ⟶ 10.655
   Then db value = $\frac{1}{2}\times 10.655 = 5.328$
   Therefore a power ratio of 3.41 = 5.33 db
4. **To convert db to power ratios.** Use the table as directed for 2 above and double the result.

E or I Ratio	.00	.01	.02	.03	.04	.05	.06	.07	.08	.09
1.0	.000	.086	.172	.257	.341	.424	.500	.588	.668	.749
1.1	.828	.906	.984	1.062	1.138	1.214	1.289	1.364	1.438	1.511
1.2	1.584	1.656	1.727	1.798	1.868	1.938	2.007	2.076	2.144	2.212
1.3	2.279	2.345	2.411	2.477	2.542	2.607	2.671	2.734	2.798	2.860
1.4	2.923	2.984	3.046	3.107	3.167	3.227	3.287	3.346	3.405	3.464
1.5	3.522	3.580	3.637	3.694	3.750	3.807	3.862	3.918	3.973	4.028
1.6	4.082	4.137	4.190	4.244	4.297	4.350	4.402	4.454	4.506	4.558
1.7	4.609	4.660	4.711	4.761	4.811	4.861	4.910	4.959	5.008	5.057
1.8	5.105	5.154	5.201	5.249	5.296	5.343	5.390	5.437	5.483	5.529
1.9	5.575	5.621	5.666	5.711	5.756	5.801	5.845	5.889	5.933	5.977
2.0	6.021	6.064	6.107	6.150	6.193	6.235	6.277	6.319	6.361	6.403
2.1	6.444	6.486	6.527	6.568	6.608	6.649	6.689	6.729	6.769	6.809
2.2	6.848	6.888	6.927	6.966	7.008	7.044	7.082	7.121	7.159	7.197
2.3	7.235	7.272	7.310	7.347	7.384	7.421	7.458	7.495	7.532	7.568
2.4	7.604	7.640	7.676	7.712	7.748	7.783	7.819	7.854	7.889	7.924
2.5	7.959	7.993	8.028	8.062	8.097	8.131	8.165	8.189	8.232	8.266
2.6	8.299	8.333	8.366	8.399	8.432	8.465	8.498	8.530	8.563	8.595
2.7	8.627	8.659	8.691	8.723	8.755	8.787	8.818	8.850	8.881	8.912
2.8	8.943	8.974	9.005	9.036	9.066	9.097	9.127	9.158	9.188	9.218
2.9	9.248	9.278	9.308	9.337	9.367	9.396	9.426	9.455	9.484	9.513
3.0	9.542	9.571	9.600	9.629	9.657	9.686	9.714	9.743	9.771	9.799
3.1	9.827	9.855	9.883	9.911	9.939	9.966	9.994	10.021	10.049	10.076
3.2	10.103	10.130	10.157	10.184	10.211	10.238	10.264	10.291	10.317	10.344
3.3	10.370	10.397	10.423	10.449	10.475	10.501	10.527	10.553	10.578	10.604
3.4	10.630	10.655	10.681	10.706	10.731	10.756	10.782	10.807	10.832	10.857
3.5	10.881	10.906	10.931	10.955	10.980	11.005	11.029	11.053	11.078	11.102
3.6	11.126	11.150	11.174	11.198	11.222	11.246	11.270	11.293	11.317	11.341
3.7	11.364	11.387	11.411	11.434	11.457	11.481	11.504	11.527	11.550	11.573
3.8	11.596	11.618	11.641	11.664	11.687	11.709	11.732	11.754	11.777	11.799
3.9	11.821	11.844	11.866	11.888	11.910	11.932	11.954	11.976	11.998	12.019
4.0	12.041	12.063	12.085	12.106	12.128	12.149	12.171	12.192	12.213	12.234
4.1	12.256	12.277	12.298	12.319	12.340	12.361	12.382	12.403	12.424	12.444
4.2	12.465	12.486	12.506	12.527	12.547	12.568	12.588	12.609	12.629	12.649
4.3	12.669	12.690	12.710	12.730	12.750	12.770	12.790	12.810	12.829	12.849
4.4	12.869	12.889	12.908	12.928	12.948	12.967	12.987	13.006	13.026	13.045
4.5	13.064	13.084	13.103	13.122	13.141	13.160	13.179	13.198	13.217	13.236
4.6	13.255	13.274	13.293	13.312	13.330	13.349	13.368	13.386	13.405	13.423
4.7	13.442	13.460	13.479	13.497	13.516	13.534	13.552	13.570	13.589	13.607
4.8	13.625	13.643	13.661	13.679	13.697	13.715	13.733	13.751	13.768	13.786
4.9	13.804	13.822	13.839	13.857	13.875	13.892	13.910	13.927	13.945	13.962
5.0	13.979	13.997	14.014	14.031	14.049	14.066	14.083	14.100	14.117	14.134
5.1	14.151	14.168	14.185	14.202	14.219	14.236	14.253	14.270	14.287	14.303
5.2	14.320	14.337	14.353	14.370	14.387	14.403	14.420	14.436	14.453	14.469
5.3	14.486	14.502	14.518	14.535	14.551	14.567	14.583	14.599	14.616	14.632
5.4	14.648	14.664	14.680	14.696	14.712	14.728	14.744	14.760	14.776	14.791
5.5	14.807	14.823	14.839	14.855	14.870	14.886	14.902	14.917	14.933	14.948
5.6	14.964	14.979	14.995	15.010	15.026	15.041	15.056	15.072	15.087	15.102
5.7	15.117	15.133	15.148	15.163	15.178	15.193	15.208	15.224	15.239	15.254
5.8	15.269	15.284	15.298	15.313	15.328	15.343	15.358	15.373	15.388	15.402
5.9	15.417	15.432	15.446	15.461	15.476	15.490	15.505	15.519	15.534	15.549

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